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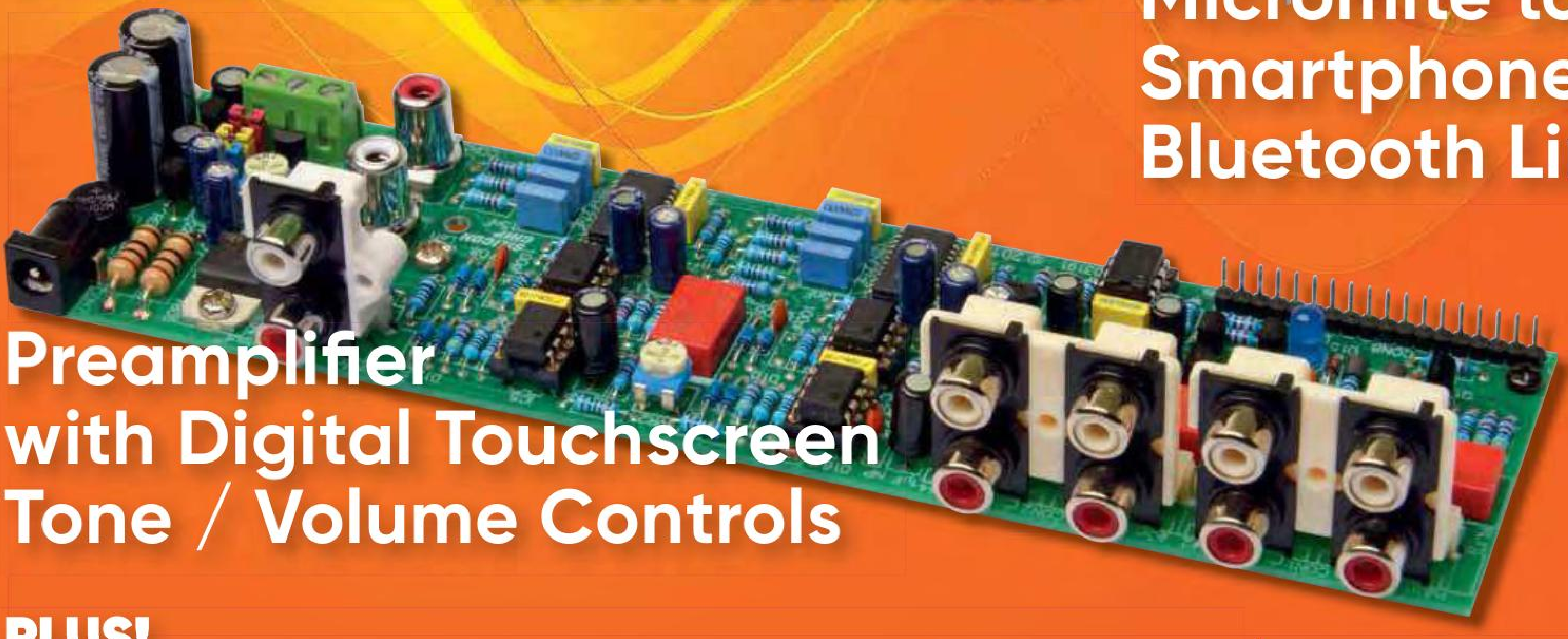

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by LOUIS MEULSTEE

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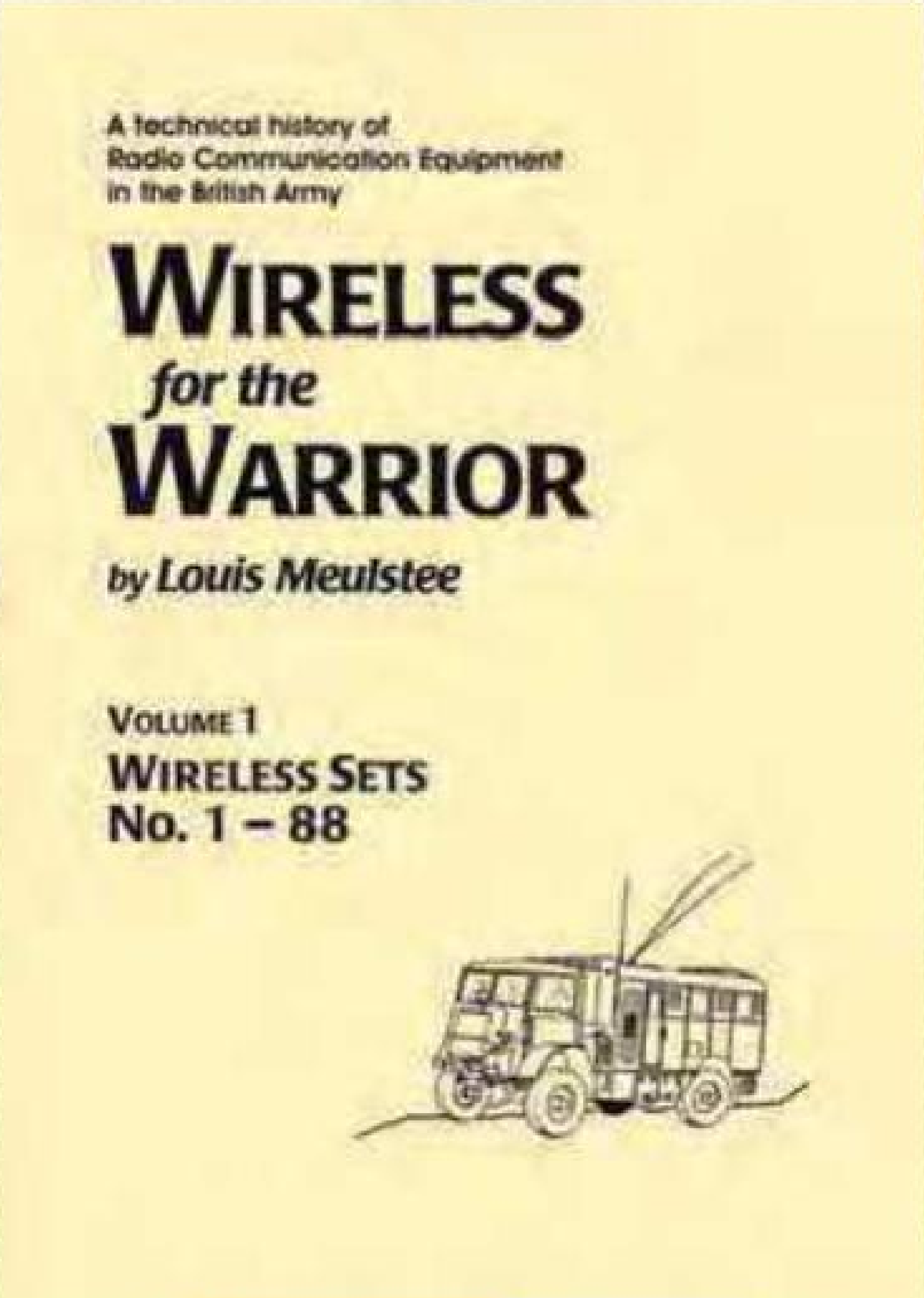

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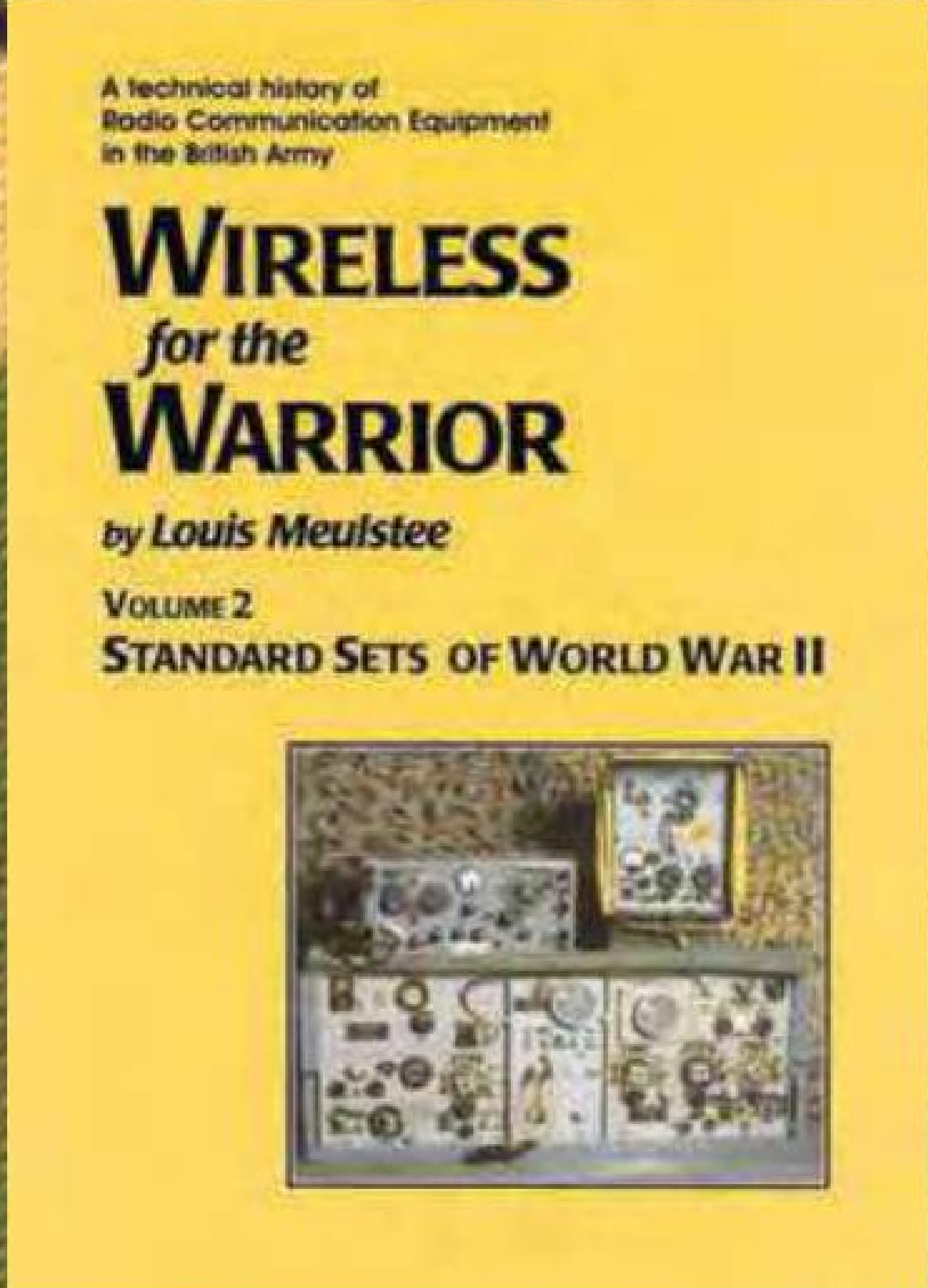

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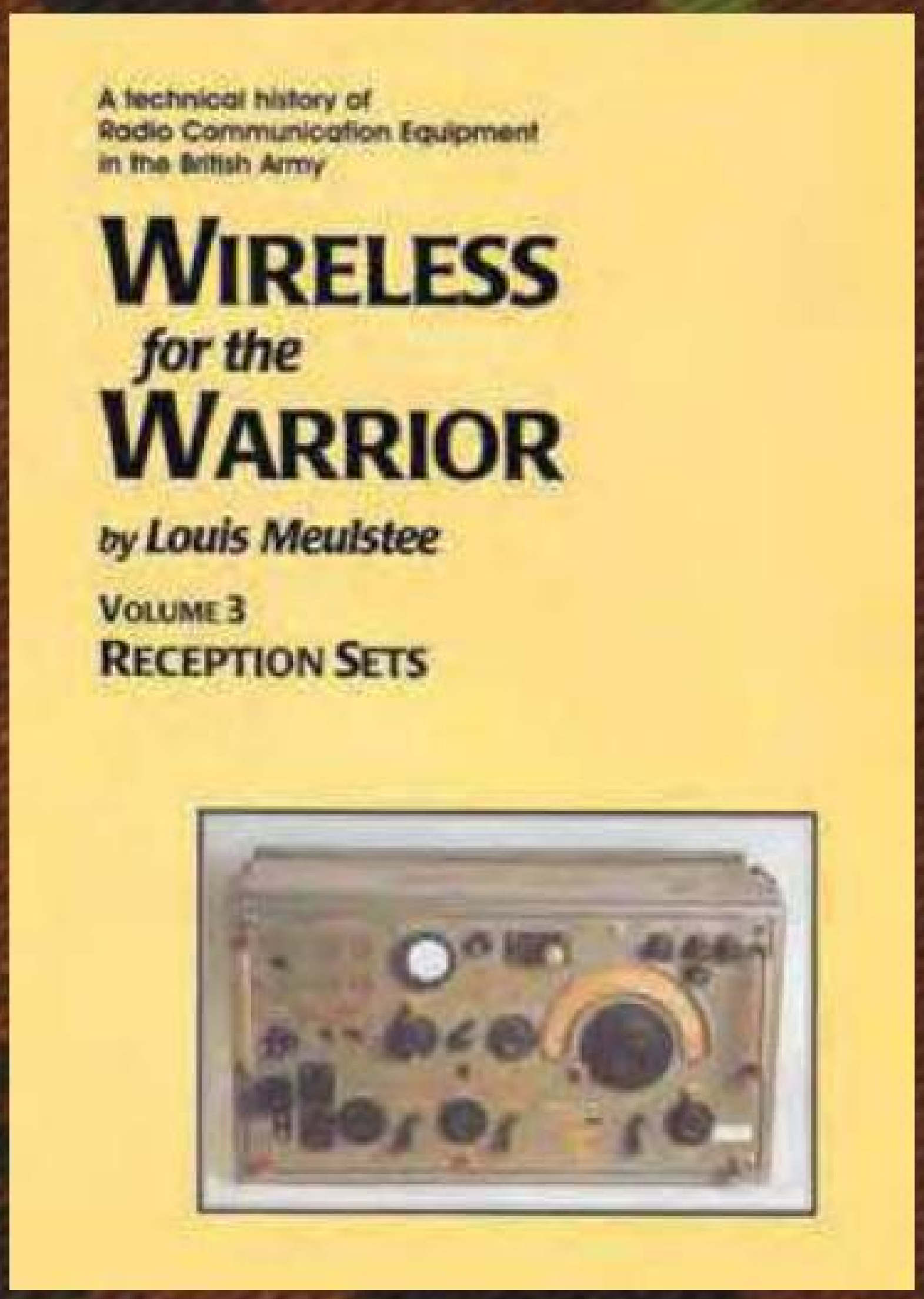

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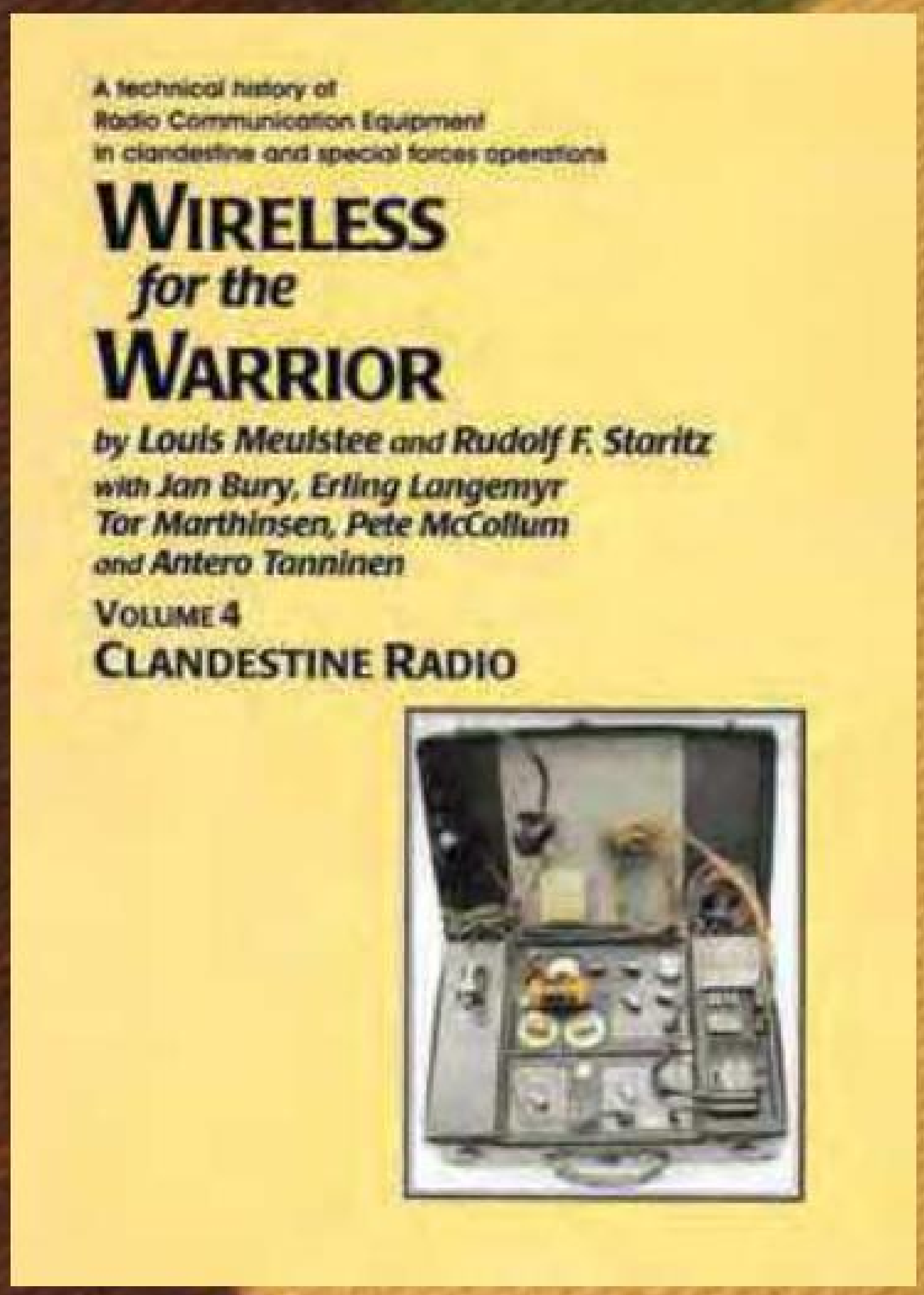

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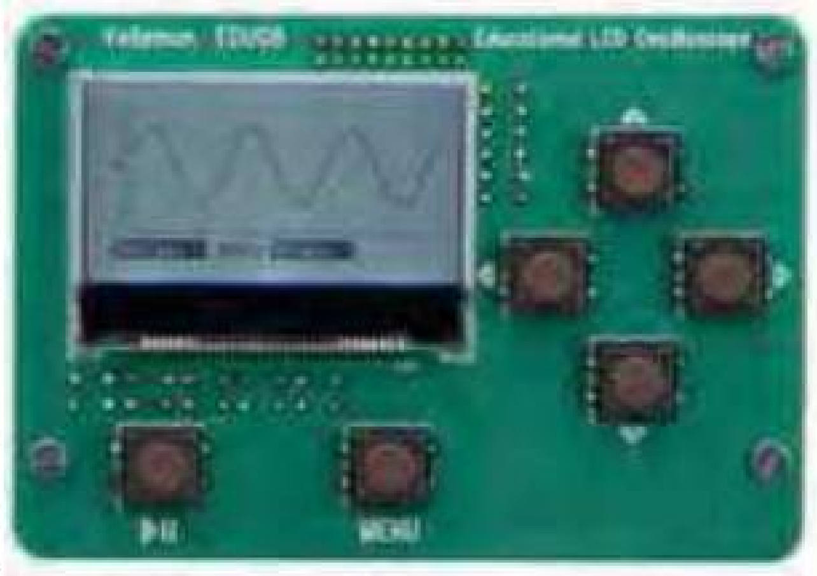

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All reasonable precautions are taken to ensure that the advice and data given to readers is reliable. We cannot, however, guarantee it and we cannot accept legal responsibility for it.

A number of projects and circuits published in *Practical Electronics* employ voltages that can be lethal. You should not build, test, modify or renovate any item of mains-powered equipment unless you fully understand the safety aspects involved and you use an RCD (GFCI) adaptor.

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Digital preamp

There are two things we know *PE* readers like – you ask for them regularly – Hi-Fi projects and coverage of digital control systems and techniques. So, it's surprising how rarely we put these two topics together. This month, we buck that trend with the start of a two-month project that combines the fidelity of an analogue design with the flexibility of digital control and a touch-screen interface. It includes Baxandall tone controls and uses a *Micromite BackPack*, so despite being low in cost it is high in features. We think you'll love it!

The power of batteries… and a warning

I wrote last month about the importance of the new generation of battery technologies and the related issue of correct charging. So, it is only right that this month Alan Winstanley reports on what can go wrong if a Li-ion battery is misused, or fails through poor design, overuse or damage. The resulting pyrotechnics can be spectacular for all the wrong reasons. Like many of you, I have a variety of DIY tools that run off a common battery system. The official branded versions of the rechargeable Li-ion batteries that power these tools are pricey, but it is becoming ever clearer that low-cost, third-party bargains could be a very foolish purchase. This is one area where it definitely pays to think in the long term and to ask yourself what am I really saving with cheap batteries, what am I risking and how would I put out a battery fire? Do you have a fire detector, fire blanket and a fire extinguisher? They aren't expensive, but if you need one in a hurry there is no substitute.

Web issues and solutions

We have had a few web issues this summer and we really appreciate the patience of those of you who have been affected. The good news is the PDF subscription website has been completely rebuilt and now runs smoothly and safely.

For digital subscribers there is more good news. When the PDF subscription system was launched it made sense to include digital security features, but 10 years on all they do is annoy legitimate purchasers and provide very little protection to the publisher. Therefore, we have removed the irritating watermarking and PDF password protection.

Matt Pulzer
Publisher

What's in a name?

Techno Talk

Mark Nelson

This month we check out various names of gadgetry encountered in everyday electronics. Some of them are informative, others silly or even a bit dated (like me!) — but still vaguely amusing in their own way.

Let's kick off with 'flower power', dating from the swinging sixties, when we all wore sandals and cheesecloth shirts, with flowers in our hair. Hobby electronics was not a dirt-cheap hobby back then, but transistors were just beginning to become affordable (especially if you bought the cheaper but slightly out-of-spec ones). Silicon devices were far too expensive for most folk, so we settled for flower power (in other words germanium) devices. Because many people were not very good at spelling, they wrote 'geranium' instead, hence the flower power nickname. Yes, hilariously amusing.

In those far-off days, transistor sets were still in the process of ousting the older valve/tube radios. Manufacturers used the trendy term 'solid state' to promote the new generation of transistorised products, prompting electronics hobbyists to invent the name 'hollow state' for sets that still used round glass things that ran hot.

Culpable and colourful capacitors

Of course, anything electronic that ran hot was prone to drifting and general unreliability. Many components could not tolerate excessive heat, which is probably how electrolytic capacitors acquired their nickname of 'smoothing bombs' (particularly the ones with ominously bulging top seals). Like most electronic components, capacitors use smoke as their active ingredient! If they are overheated or overloaded, they explode and release the smoke that up till then had powered them. This is why you should always conduct a 'smoke test' to check whether a newly built or repaired gadget will run for several hours without liberating any smoke.

Some folk will tell you that smoke testing is not a silly name invented by electronicists, but rather an entirely rational term borrowed from the plumbing trade. After new pipework has been installed, smoke is created artificially, blown into the tubing, thereby enabling an observer to check the pipes and make sure no smoke is leaking. There may be some truth in this assertion because an alternative name for this exercise is a 'soak test', in which newly constructed or repaired equipment is observed under stressful condition over an extended period. Burst pipes would be a clear indication of failure. But we are drifting off-topic, so let's get back to dodgy capacitors.

Highly suspect are the capacitors nicknamed 'humbugs', 'black beauties' and 'bumblebees', which were made in the 1950s – all three being of similar construction. They are notorious for drifting in value or failing to block DC properly, thus contributing to overheated anodes (by putting a false bias on the next stage of amplification) and early valve death in general. Their weak point was a Bakelite resin shell that all too often cracked or crumbled, allowing moisture to enter and degrade the paper dielectric.

Humbugs looked strikingly similar to the old-fashioned mint humbug sweet, while bumblebees took their name from the coloured stripes used to indicate their value. Most electronicists replace these caps without bothering to even test them. Far more desirable, both in looks and reliability, were and are the Mullard/Philips C280 miniature metallised metal film capacitors, dating from the mid-1960s. Known universally by their 'tropical fish' nickname, they remain highly regarded for their immunity to moisture and temperature drift.

Same item, different name

Let's take the humble phono connector – not to be confused with a *phone* connector of course. Why does such a commonplace item have at least three names in the English language? And why phono? Answering the second question first, it's a throwback from when owners of radio receivers (wireless sets) could for the first time connect a record deck or electric phonograph. On the rear of these upmarket radios was a small coaxial connector marked 'phono' for plugging in the connecting cable. Presumably, this feature originated in the US, where gramophones were called phonographs.

Wikipedia confirms this and adds that the connector was launched in 1937 by the RCA company, which explains why most Americans calls these plugs 'RCA connectors'. On the other hand, many people call them 'Cinch connectors,' a term widespread in continental Europe. Historically, the British and US Cinch companies made all manner of connectors, but whether they also made phono connectors I cannot say at this moment.

Another classic design of connector with many applications in electronics is the barrel connector, which is a small coaxial connector used for hooking up a low-voltage power source to both consumer electronic devices and a host of industrial products. The (presumably American) author of the Wikipedia page accords it the alternative names of 'concentric barrel connector' and 'tip connector', but sadly omits the name 'ROKA connector', which is used widely in Europe. Why sadly? Because the design of this connector that is now used all over the world originated in Germany and derives from the initials of the Robert Karst electronics company.

While we are on this subject, it's worth noting that conventionally these connectors are centre-positive, but this applies only in consumer electronics. In the US, the reverse applies on professional and industrial equipment, the centre pin being negative. If a gadget using these power connectors appears to be dead even though the power supply is known to be good, the reversed method of wiring may be the cause. Note that the pin of these connectors can measure either 2.1mm or 2.5mm in diameter (2.1mm is more commonplace). A 2.5mm power connector plugged into a device equipped with a 2.1mm pin will not make contact. I once discarded a perfectly good 'wall wart' power supply after failing to spot this stupid mistake!

One final case of pluggery with multiple names is the mini jack connector, also known as a 3.5mm or 1/8-inch plug or socket. The proper name for the socket is a 'jack' and the corresponding 'plug' is the proper name for what some folk confusingly call a 'jack plug'.

Win a Microchip PIC32CM LS60 Curiosity Pro Evaluation Kit

Practical Electronics is offering its readers the chance to win a Microchip PIC32CM LS60 Curiosity Pro Evaluation Kit (EV76R77A) – and even if you don't win, receive a 15%-off voucher, plus free shipping for one of these products.

The PIC32CM LS60 Curiosity Pro kit is ideal for evaluating and prototyping with the secure and ultra-low-power PIC32CM LS60 Arm Cortex-M23 based microcontrollers. The MCU integrates the TrustFLEX ECC608 secure subsystem and Arm TrustZone Technology in one package. Additionally, it features enhanced peripheral touch controller (PTC) and smart analogue, such as op amps, ADC, DAC and analogue comparators.

The PIC32CM LS60, which integrates Microchip's Trust Platform secure subsystem, makes it easier to develop end products using one microcontroller rather than two or more semiconductor chips. Now, designers can go to one trusted source to find a 32-bit MCU that is designed to protect products and the end user from remote or physical attacks on their smart home devices, smartphone or tablet accessories, portable medical devices, wearables, connected appliances and industrial robots.

It has an Arduino Uno connector, mikroBUSTM socket and extension headers to expand the development. The kit includes an on-board Embedded Debugger, eliminating the need for external tools to program or debug. The PIC32CM LS60 is compatible with the MPLAB Data Visualizer and Power Debugger tools that can be used to monitor, analyse and fine tune power consumption numbers in real-time. This makes it easier for customers to develop low-power applications and release extended-battery-life products to market faster.

Worth $77.00 (approx £62.50) each

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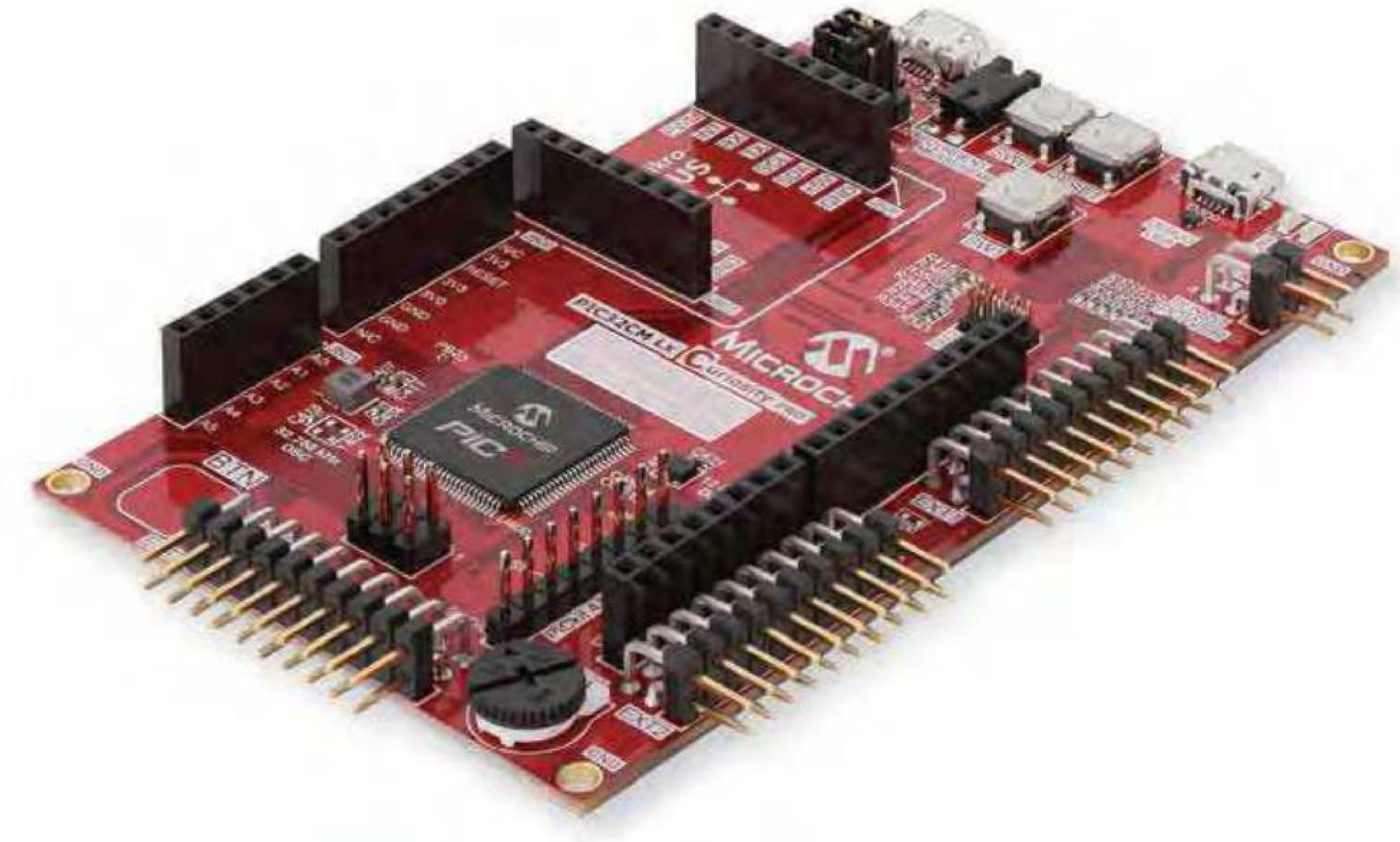

Microchip PIC32CM LS60 Curiosity Pro Evaluation Kit

How to enter

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Closing date

The closing date for this offer is 31 August 2022.

May 2022 winner

Orlando Matos

He won a Microchip Explorer 8 Development Kit

The **Fox** Report

Barry Fox's technology column

How accurate is smart meter advertising?

This winter – or probably well before then – homes with 'smart' meters will discover an unhappy truth. When they took up the offer of a free meter, they gave the gas and electric supply companies freedom to cut off their power remotely, at the flick of a switch (or more likely, the click of a mouse).

Unsurprisingly, this doesn't feature in the power companies' advertising campaign for smart meters, co-ordinated by a company called Smart Energy GB. All the promotional emphasis is on reducing bills and saving the planet with wind turbines and solar power.

Of course, it's good that customers who could pay their bills but persistently choose not to pay, and try to deny entry to cut-off engineers, will now get their just desserts. But honest victims of unhappy circumstance risk being stuck on Kafkaesque help lines trying to get re-connected. Some may feel they were misled by the power companies' promotions. Questions will then be asked.

Legal, decent, honest and truthful?

So, this month, I will distil the essence of countless hours of pressuring the Advertising Standards Authority over the adverts for smart meters it has blessed. I have tried – so far in vain – to interest officialdom. If any *bona fide* body, with the clout to investigate, now shows genuine interest and concern, I will provide access to chapter and verse documentation. If any reader wants to circulate a copy of this article, for instance to their MP, feel free. I can be contacted via my website: **www.tekkiepix.com**

In November 2021 I complained to the ASA about adverts (eg, *Evening Standard*, 15 November 2021) which trumpeted that 'More wind power is a smart thing. Smart meters are helping modernise Britain's energy system, so it can use more renewables and solar power'.

I baulked at the 'so it can use' claim. How does the use of smart meters in homes enable wind turbines and solar farms? Smart meters enable the energy companies to save money on human meter-readers, and remotely disconnect supplies. Without evidence or explanation, 'so it can use' seems to be a complete *non sequitur*.

Silence isn't golden

The ASA acknowledged receipt of my complaint but then fell silent. This is because the ASA has recently 'convenienced' itself – if no one else – by changing its complaint-handling procedure. Anyone complaining about an advert now gets a standard form acknowledgement, but then hears nothing further unless the ASA decides to follow through on the complaint. The complainant is, therefore, left dangling with no information on whether or not their complaint is being followed through, and if not, why not. There is no guidance given on how long a silence signals a considered decision not to act.

The ASA acknowledged, 18 months ago, that my, 'suggestion that we should indicate a likely time frame within our acknowledgement email is something that struck a chord immediately and we will look to implement something like this'.

'It's a very helpful suggestion, thank you,' said a senior spokeswoman. But the last time I looked, the ASA system was still leaving complainants dangling.

When I pushed the ASA hard on the lack of action on my smart meter complaint, a senior ASA spokesman explained that it, 'raised an issue we have already considered, and therefore did not need to pursue further.'

'The ASA has previously investigated environmental claims by Smart Energy GB' the ASA spokesman explained, and pointed to a ruling from September 2019: **https://bit.ly/pe-sep22-asa1**

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This said: ‘The ASA Council considered evidence supplied by the advertiser and understood that the data captured and reported by smart meters was intended to help the national grid to anticipate demand and plan energy provision more effectively, with less reliance on fossil fuels. It therefore concluded that the ads did not break the rules’.

In early 2022 I reminded the ASA that adverts from Smart Energy GB (*Evening Standard*, 24 January 2022) were still conflating the fitting of smart home meters with renewable energy sourcing. I warned that the burgeoning energy crisis and likely rash of remote cut-offs could trigger questions on the ASA’s failure to curb the questionable claims.

On the one hand…

On the face of things, the ASA is not acting on new complaints about smart meter adverts, because it has previously considered and allowed previous adverts. But the older averts were far more cautiously worded than the new adverts. And it is devilishly hard to decode the ASA’s policy on this.

Chief Executive Guy Parker points to one ruling, while an ASA spokesman points to different ruling. Guy Parker cites a March 2019 ruling on smart meters as, ‘partly upholding against Smart Energy GB.’ – see: **https://bit.ly/pe-sep22-asa2**

This ruling flowed from adverts run in mid-2018. 14 people objected to claims that installing gas and electric smart meters in homes saved energy nationwide because people immediately saw how much they were spending.

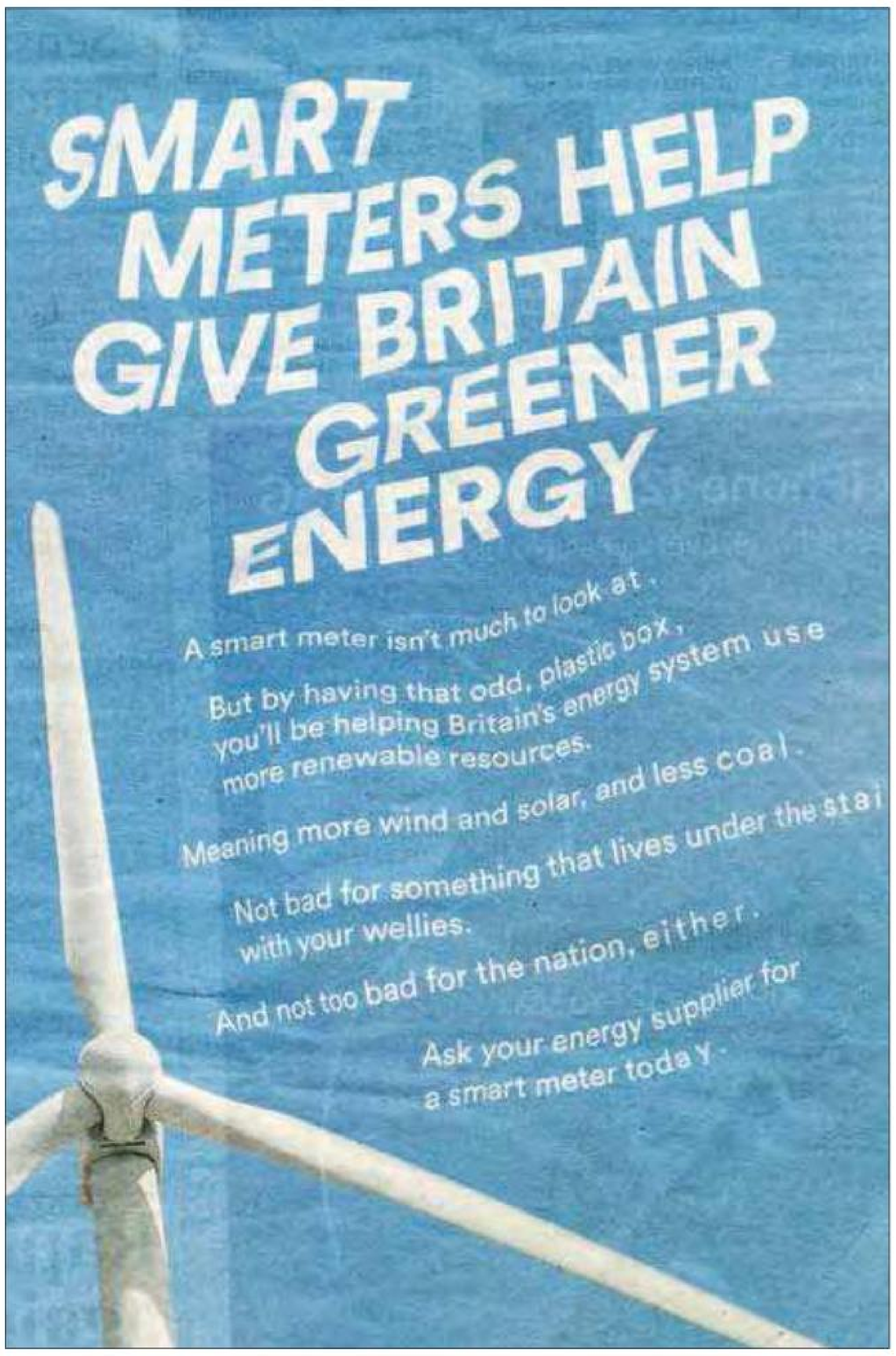
Few people with a meter would argue with this. The meter readout shows red as an oven gets hot, a kettle boils or when a room heater heats. So, the householder becomes more careful.

What the ASA didn’t like was a suggestion that the meter itself was saving energy, and the complaint was party upheld. But there had been no attempt at linking meters with wind or solar power. Thus, the March 2019 ruling is a red herring.

Separately, an official spokesman for ASA points to a ruling from September 2019 which condoned adverts run in April 2019 that tied meters to cleaner air. But these adverts included a very significant qualification: ‘Smart meters can’t solve climate change on their own, but with the smarter, more energy efficient grid they help to create, they’re a start’. So the complaint was not upheld – see: **https://bit.ly/pe-sep22-asa1**

I have struggled to find similar acknowledgments in more recent adverts which the ASA hasn’t investigated. A senior ASA spokeswoman says it, ‘was not a requirement, as a result of the previous investigation…. to include this as qualifying information in their ads’.

This could very well explain why such an honest qualification does not appear in more recent promotions for smart meters.



SMART METERS HELP GIVE BRITAIN GREENER ENERGY

A smart meter isn’t much to look at.

But by having that odd, plastic box, you’ll be helping Britain’s energy system use more renewable resources.

Meaning more wind and solar, and less coal.

Not bad for something that lives under the stai with your wellies.

And not too bad for the nation, either.

Ask your energy supplier for a smart meter today.

Rising complaints

More heavy pushing led the ASA to confirm that since 2014, it has received a total of 180 complaints relating to 21 discrete cases which relate to the role of smart energy meters in the supply of renewable energy. Only one (the one from September 2019 which contained the frank admission that ‘Smart meters can’t solve climate change on their own, but with the smarter, more energy efficient grid they help to create, they’re a start’) was considered to be suitable for formal investigation – but ‘not upheld’.

‘Other cases’ says the ASA ‘were either informally resolved, not investigated after initial assessment because they did not raise any actionable issues or were out of the ASAs remit.’ Presumably mine fell into this hole.

For completeness, I also asked the ASA several times whether the Mark Lund named as Chairman of Smart Energy GB (the company behind the smart meter adverts) was the same Mark Lund named as Chairman of the ASA’s funding body ASBOF (The Advertising Standards Board of Finance). I made and make no suggestion of any wrong-doing, I just want to get the optics clear and the facts straight.

ASA Chief Executive Guy Parker has now clarified the optics on Mark Lund’s dual role: ‘Mark chairs ASBOF and BASBOF our funding boards. The ASA operates independently and at arms’ length from ASBOF and BASBOF which are responsible for collecting the industry levy/fees from advertisers, online platforms, etc … neither Mark nor any other Board members of ASBOF and BASBOF play any role in the ASA’s decision-making when it comes to apply(ing) the UK Advertising Codes.’

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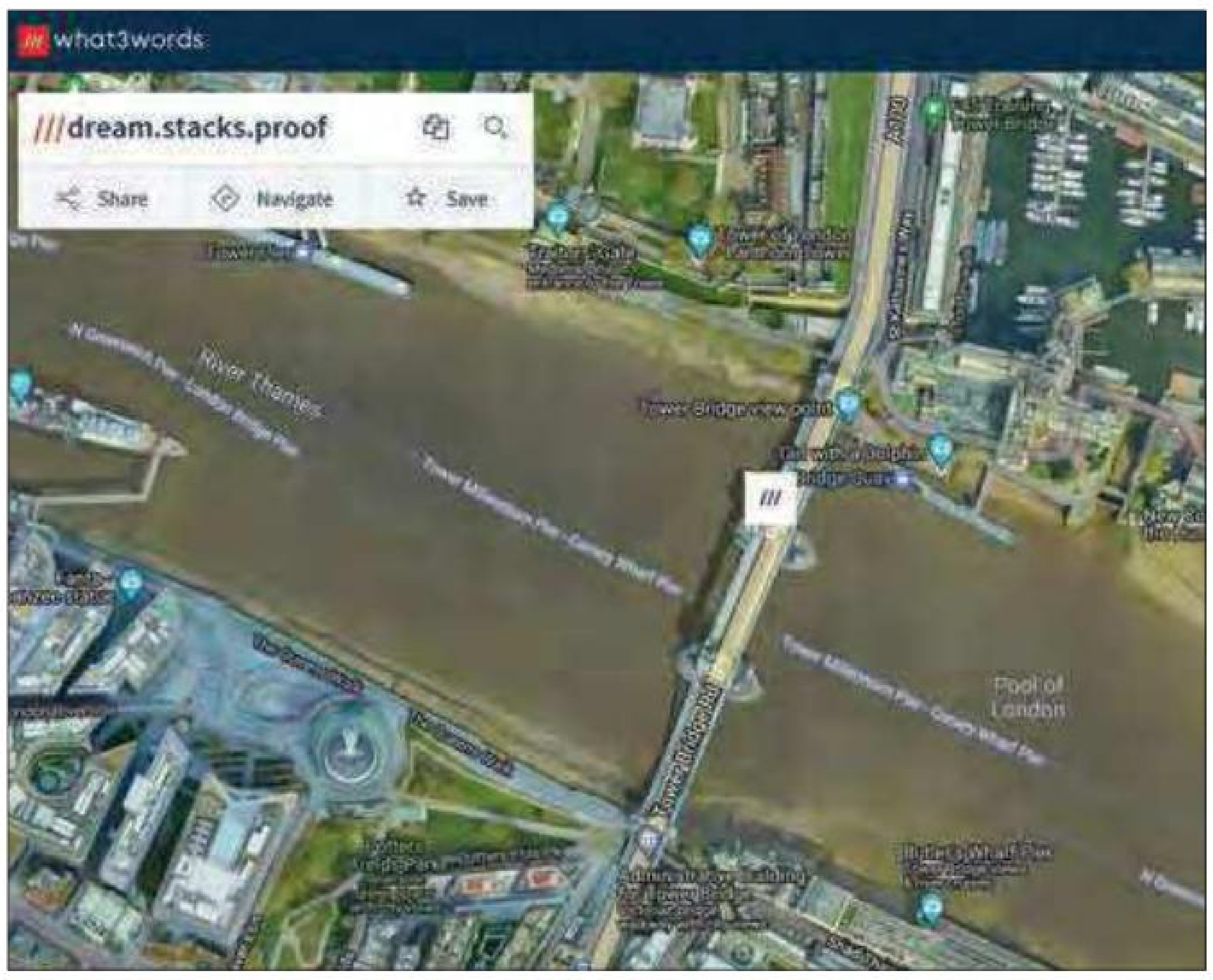
Net Work

Alan Winstanley

This month's roundup of technological trends investigates the worrying problems of lithium battery fires, ponders whether (and how) to upgrade to Windows 11 and spells out the things What3Words can do for you.

According to the 16th century proverb, 'Three may keep a secret, if two of them are away'. However, the use of three words is central to the idea behind London-based What3Words, a geolocation service that enables users to pinpoint an address by using three easily digestible words. The firm has divided the globe into a grid of 3-metre squares, which they decided was the resolution needed to enable an area the size of an entrance or a car parking space to be pinpointed accurately. To map the entire planet (seas, oceans, lakes and all) would need 57 *trillion* addresses, which they calculated could be described using combinations of three words. The use of three words also gave them seven trillion addresses to spare, as two words would only have been enough for 1.6 billion squares. The system is future-proof, because even if buildings or addresses change over time, the same three words will always describe that same geographical location on the globe.

Many safety benefits come with using What3Words, as the system makes it easy to describe one's precise location in an emergency. In the UK, 80% of emergency services can now recognise What3Word phrases and it is catching on overseas, including with the Canadian police. In some cases, services can text a 'FindMe' link to an incoming caller which will generate the caller's What3Words location. Businesses can use What3Words to help with logistics or the navigation of company vehicles. It is also finding its way into vehicles, with Subaru USA, DPD Parcels UK and Jaguar Land Rover adding What3Words to supplement navigation systems. Across India, street names are often duplicated, addresses can be confusing, and roadsides and many rural areas don't have addresses at all. So Indian vehicle maker Mahindra announced last month it is integrating What3Words with Amazon's Alexa in its all-new Scorpio-N SUV.



What3Words uses easily digestible words to describe any location on the planet. London's Tower Bridge is described here in the desktop version.

Way to go: Indian vehicle maker Mahindra is incorporating What3words with Alexa into its new Scorpio-N SUV.

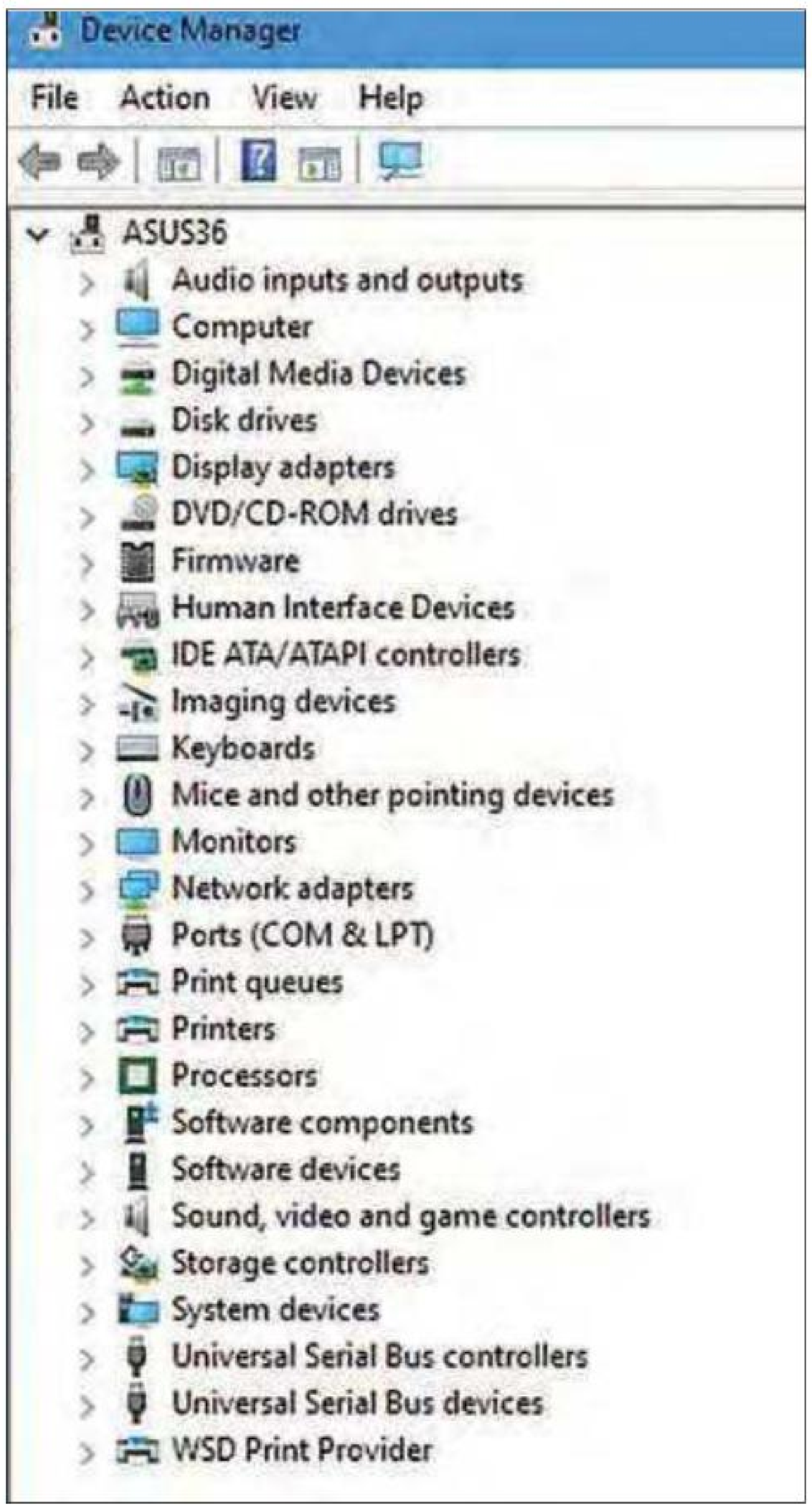
Multi-language integration

To make a start with What3Words, an app is available for mobile use that will generate the three-word string, but it can also be used offline with GPS, which will provide the three-word location without reference to a map (see the guide at: **https://youtu.be/_JBwRxjCq-k**). If you have no mobile phone signal, you will need another means (radio, landline) to communicate those three words in order to summon any services. The website **https://what3words.com** enables desktop users to zoom in to any address on a map and the three words will be displayed. Both map and satellite layers are available, and a typical house might cover ten or twelve squares, so you can almost pinpoint your living room using What3Words. Looking ahead, the company also states that it's the only addressing system optimised for voice, making it the ideal location system for the voice-enabled solutions of the future. We can expect to see more voice recognition and Alexa integration in due course. It has been made available in 51 languages.

A word or three about logins

The use of three whole words is also suggested as a way of managing personal passwords. It's becoming increasingly pointless to use random character strings like `3hroY7h0#_67`, which is a login that can't be remembered and will probably end up written on a Post-It note somewhere. A nonsensical passphrase could be created by stringing three words together, such as `poachedeggssalt22`. At least they are more human-friendly and usable, and no more likely to be hacked than any other combination of characters. Britain's National Cyber Security Centre (NCSC) explains the rationale of using three random words this way, at: **https://bit.ly/pe-sep22-ncsc**

However, the use of biometrics and two-factor authentication (2FA) is increasingly replacing the need for passwords altogether. In China, facial snapshots are increasingly used as a means of payment verification: Facial-Recognition Payment (FRP) or 'Scan the face to pay' systems can look up your picture on their database and complete


Device Manager

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- Ports (COM & LPT)
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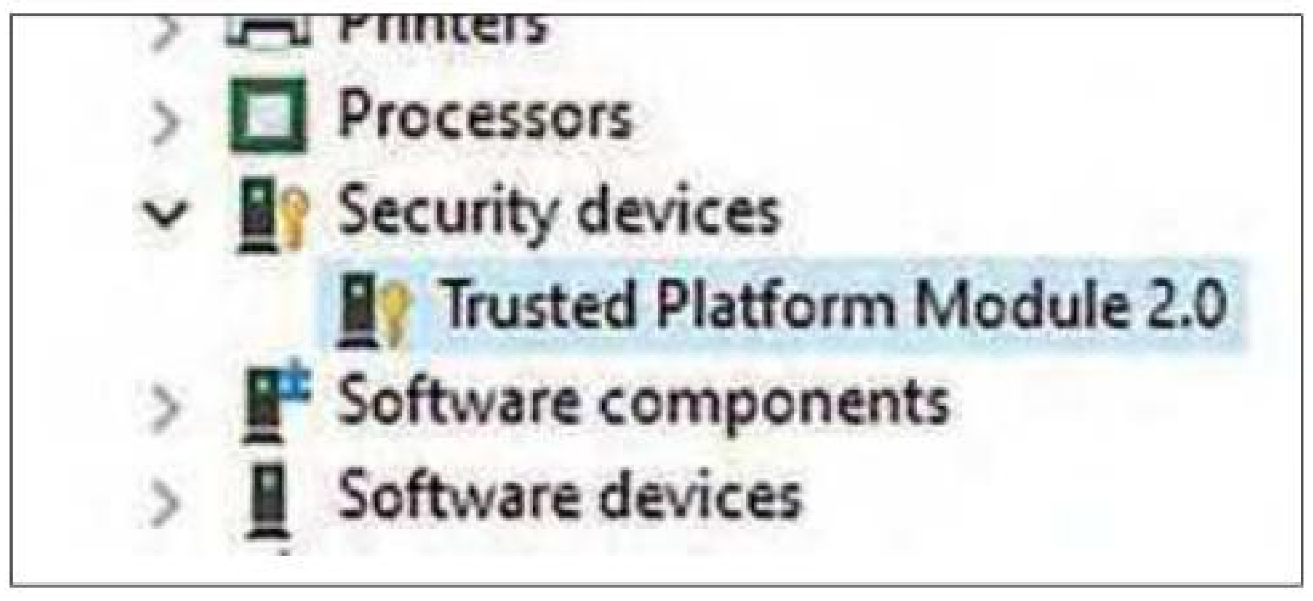
- Printers
- Processors
- Software components
- Software devices
- Sound, video and game controllers
- Storage controllers
- System devices
- Universal Serial Bus controllers
- Universal Serial Bus devices
- WSD Print Provider

Before (above/right) and after (below) a BIOS update, which 'enabled' an onboard TPM2.0 module, allowing a Windows 11 upgrade.


- Printers
- Processors
- Security devices
  - Trusted Platform Module 2.0
- Software components
- Software devices

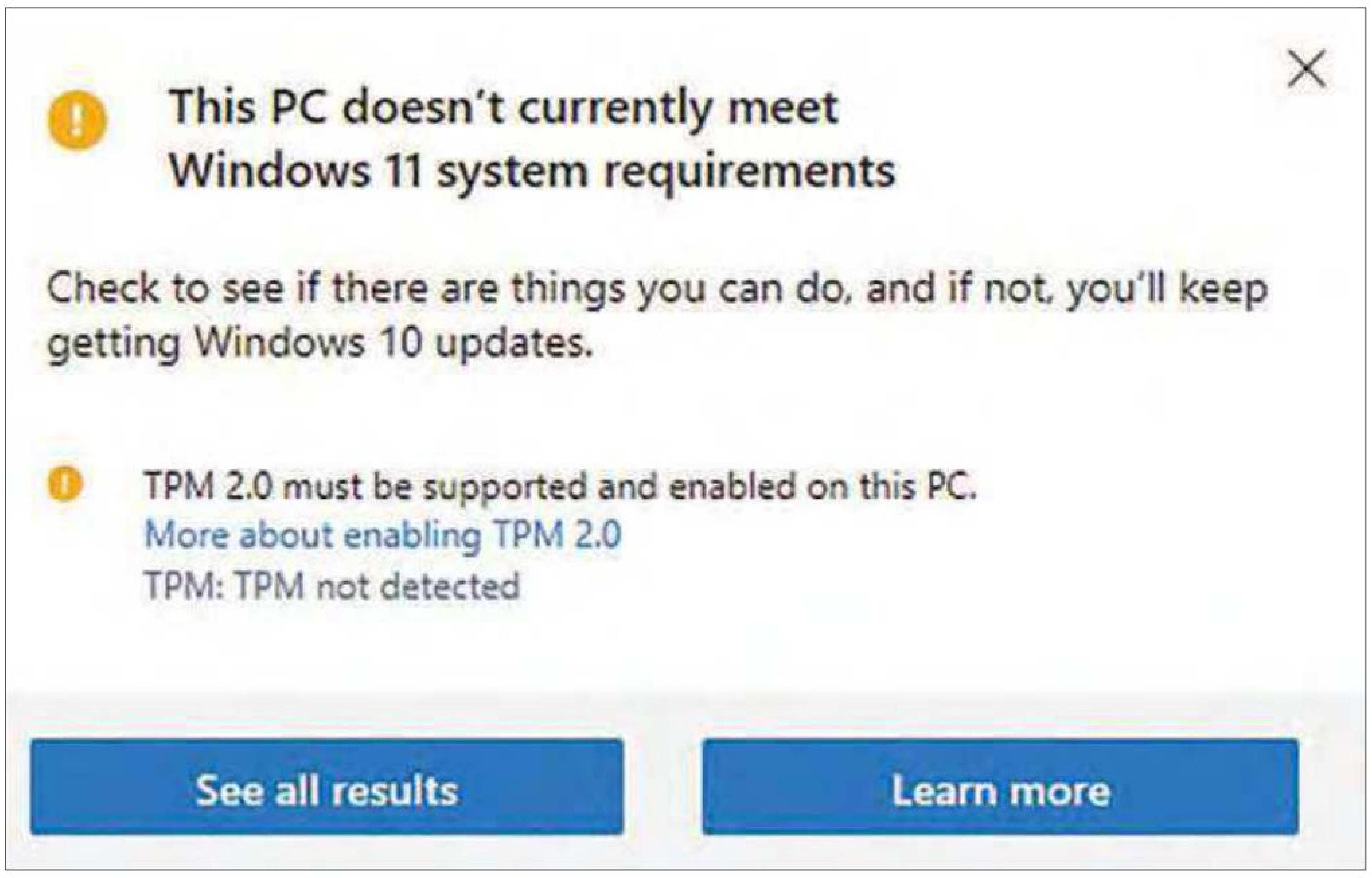
a transaction within 10 seconds. (The Nielsen Norman Group published a case study of FRP two years ago at: **https://bit.ly/pe-sep22-frp**)

Managing logins

As a regular web user, I've relied on Roboform for many years to manage several hundred web logins. Roboform Everywhere for Windows, Mac and mobile is cloud-based and it wakes up to autofill web browser logins with ease; plus, it stores 'Safe Notes' securely as well. Roboform uses an optional master password and is Windows Hello compatible, which calls for a PIN, fingerprint or webcam scan. (For desktop PC users, USB fingerprint readers that are Windows Hello compatible cost £20-25 on eBay.) Roboform Everywhere starts at $17.90 a year and a free version is worth trying. Readers can visit **Roboform.com** for more information.

Windows 11: Check your TPM

Regular *Net Work* readers may recall my piece in September 2021 about the then-forthcoming version of Windows


This PC doesn't currently meet Windows 11 system requirements

Check to see if there are things you can do, and if not, you'll keep getting Windows 10 updates.

TPM 2.0 must be supported and enabled on this PC.
More about enabling TPM 2.0
TPM: TPM not detected

11 and its minimum system requirements. One obstacle to upgrading older PCs from Windows 10 can be the lack of a Trusted Platform Module (TPM), which is a hard-wired chip-level cryptographic device that W11 uses to enhance data security. Whether you can upgrade to Windows 11 depends on whether your motherboard supports a TPM 2.0 module. You can check for W11 compatibility by typing 'PC Health Check' into the W10 search bar and running the app. True to form, my home-brew Asus W10 PC claimed it did *not* meet the upgrade requirements as TPM2.0 was not supported. Yet my Asus motherboard had what the makers term 'Secure Boot', meaning it should be upgradeable to Windows 11. The first port of call was the BIOS settings to see if something like Secure Boot needed enabling there: no joy. A BIOS update was then suggested, which a seasoned PC user can handle (ensuring you have older BIOS files on disk should you need to revert back to them). I duly updated the BIOS, and – *lo!* – TPM2.0 suddenly appeared, as shown in my PC's Device Manager settings.

This can kick-start the well-oiled Windows Update routine before you can even blink, but it raises the question of whether you actually *want* to upgrade. Will existing software run on Windows 11? For example, I wrote previously how Epson software for a V600 flatbed scanner was not fully compatible with W11, and Epson's dust removal imaging technology – one of the reasons people buy that scanner – would be lost (see *Net Work*, April 2022). The web is alive with complaints of W11 incompatibility. Some legacy software may also be difficult or messy to re-install, even if it works at all. So, for now, readers, I've reverted back to the older BIOS: I'm happy to sail along with a PC that stubbornly refuses to upgrade to Windows 11. A lot of useful info is at the Asus Windows 11 FAQ: **https://bit.ly/pe-sep22-asus** (it will also help users of other brands).

One born every minute

At about the same time, I also applied the 'three words' idea to one of my POP3 mail accounts. I then hit another annoyance: my PC kept sounding a notification (the Windows 'Balloon' sound), every minute, on the minute. It would not be silenced! I tried everything under the sun to locate the cause of this annoying beep. Maybe it was caused by my BIOS downgrade? Or anti-virus software?

Running Process Monitor (see **https://bit.ly/pe-sep22-pmon**) showed Windows' inner workings in real-time (no clue there), and Windows' built-in Event Viewer displayed Windows logs and error messages (ditto). I almost tore the PC apart looking for the cause, before silencing the sound by turning audio off altogether. This madness went on for a day or two, before I realised that, strangely, the POP3 mail that I previewed on a tablet or smartphone did not appear on my regular PC. Then the truth dawned – I had failed to update my PC's Mailwasher POP3 settings for the new three-word login, so Mailwasher was trying to fetch mail every minute in the background before sounding the alert because the POP3 logins were wrong! Software makers Firetrust Limited have offered to add a visual warning of errors into a future release of Mailwasher.

Spontaneous com-bus-tion

Following my item on exploding electric buses that go into volcanic meltdown due to battery failures, my thanks go to regular long-time *PE* reader Godfrey Manning, who writes: 'The Paris exploding bus (*Net Work*,

A crash-damaged Tesla EV ignited three weeks later after arriving at a California scrapyard. (Sacramento Met. Fire/ Facebook)

July 2022) isn't the only example of a vehicle spontaneously combusting. One did this standing outside the garage in Potters Bar (Hertfordshire, England) a few weeks ago. The uncontainable blaze then caught several other buses nearby. The lack of injuries was only down to luck, not science. A long while ago, a battery in an emergency locator beacon caught fire in an aircraft remotely parked and empty at Heathrow Airport, London. Good thing it wasn't in the air.

'It seems to me that we're too ready to rush new technology from lab to production. It's well known in engineering (and pharmaceuticals too) that unwanted effects become apparent when production quantities are in service, phenomena that were unlikely to manifest at the small-scale testing stage. We need to stop implementing techniques just because we can and should slow down and consider what's really beneficial and even then, start with small-scale trials.'

Harking back to my 1970s school days, one day my Nuffield 'O' Level chemistry teacher was expounding the properties of the alkali metals group of elements, the family that includes lithium (Li), sodium (Na) and potassium (K). Glass-fronted cabinets lined the 'chemmy lab' and they stored a vast array of glass jars containing samples of most elements and chemical compounds for your scribe to study. Small lumps of alkali metals were stored in individual jars, submerged under oil, to prevent contact with air or moisture. One of my erstwhile chums decided to smuggle a small lump (potassium, I think) out of class, to 'have fun with it' at playtime. Afterwards, he tried flushing the piece of potassium down a school toilet: the resulting explosive reaction demonstrated how alkali metals and water really didn't get on well together!

While not wanting to be accused of scaremongering, it's worth knowing how the rapid uptake of large-scale rechargeable lithium batteries is creating a whole new raft of risks and challenges, especially for those faced with tackling problems when battery technology goes wrong. I must admit I'm starting to look at lithium batteries with a more wary eye.

Hot wheels

I found footage of the incident Godfrey highlighted on YouTube (see: **https://youtu.be/peT72jWHXfM**). These intense battery fires seem to start spontaneously, and large-scale fires seem to be virtually impossible to deal with, other than by letting them burn themselves out. In June, a written-off Tesla electric car that had been stored in a Californian scrapyard for three weeks was suddenly engulfed by fire. The local Sacramento fire crew extinguished the body shell but residual heat in the wreckage kept re-igniting gasses that were escaping from the car's battery pack. The fire raged even when the car was turned onto its side and drenched with water directly; the vast amount of water running off also risked polluting the land. 'It took a significant amount of time, water and thinking outside the box to extinguish,' said the fire chief. Eventually firefighters built a bund wall around it and filled it with water to let the car cool off. The tactic ultimately reduced the total amount of water needed, estimated at 4,500 gallons (17,000 litres), and it limited the runoff of contaminated water as well.

Tesla battery fires aren't exactly new: a year ago a news report claimed that a new Tesla Model S caught fire just a few days after delivery. The driver struggled to get out of the car (the doors wouldn't open) and he had to force his way out through the door to escape. The fire needed 'copious amounts of water' for over two hours before it was finally extinguished. The CNBC news clip is at: **https://youtu.be/t2adVW6bTkQ**

In May this year, a Tesla owner in Vancouver had to break out of his car, again by kicking the windows out, when the car suddenly stopped working and smoke filled the cabin. A report is at: **https://youtu.be/dQxm6n7SdvE**

YouTube is full of electric vehicles burning out: in the five-minute video at **https://youtu.be/Le6KNI9YsH0** showing an EV charging bay, at 0:30 seconds the sound of escaping gas (presumably from the battery pack gassing) escalates before fire breaks out. Note the intense blowtorch effect on several sides, which ultimately destroyed three vehicles.

Scary e-scooter meltdowns

The risks of lithium battery failure also apply to lesser forms of transport. British Customs authorities have long confiscated imported e-scooters that have fake safety certification or non-compliant batteries or chargers. In January, an e-scooter, stored in the

An e-scooter battery ignited without warning while stored in a community area of an apartment block in Essex. (Essex County Fire & Rescue) communal area of a block of flats in Harlow, Essex, suddenly caught fire. A passer-by took it safely outside, but the outcome could have been much worse if the e-scooter had been indoors: that's exactly what happened later on in June when a fire on the 12th floor of a block of flats in Shepherd's Bush, London, broke out and wrecked the apartment. An e-scooter had overheated and caught fire while being charged. Eight fire engines and 60 crew attended: there were no casualties, but the incident focused minds as it was less than a mile away from the Grenfell Tower block that was engulfed in 2017 when cladding caught fire, and 72 residents lost their lives.

In the Shepherd's Bush fire, a new system called '999EYE' was used that enabled callers to live-stream video straight to the Fire Brigade's 999 control centre (999 being the UK equivalent of 911). 999EYE was developed by Capita and West Midlands Fire Service and it sends a link to the caller's smartphone which starts the camera (think, having a video consultation with your GP). The streaming video helped firefighters to assess the risks that might be involved – see: **https://youtu.be/8E-DVij0km8**

In Singapore, a resident had a very lucky escape when his e-scooter suddenly exploded like a Roman candle while charging. The video – shown here: **https://youtu.be/7u_6jCVK6Us** – is shocking. The same happened to a UK e-scooter owner, see: **https://youtu.be/UqH2CJBRWP4** (language/NSFW). Finally, an ordinary lithium battery is shown erupting at: **https://youtu.be/nFgXFIa8luY** (skip to 4 mins 40 seconds – but do watch to the end).

Battery early warning signs

It should be emphasised that these are extreme cases, but it's worth knowing how to recognise some early warning signs. At consumer level, lithium batteries appear in everything from earbuds to laptops and vacuum cleaners to power drills. In some failure modes, they can short internally (a 'soft short circuit') that can be undetectable from outside. It's also not uncommon to see small Li-ion batteries that have swollen in size, due to gassing as the battery starts to age. A Chinese 'compatible' digicam battery swelled suddenly and jammed itself in the author's camera, as did a smartphone battery, and an old 18V 4Ah battery on the workbench was found to have swollen and cracked overnight, causing me to wonder whether it would short out internally and maybe ignite; not what I want in a workshop full of flammable materials (and a car). After seeing those videos, I shall scrap it immediately. You might want to buy a fire blanket and extinguisher, just in case!

Larger batteries may incorporate internal strapping and vents to help overcome swelling, but in extreme cases any damage or pressure that builds up may puncture the separators of the cells and create a chemical short circuit, risking thermal runaway and an intense fire. Swollen lithium batteries of any shape or size are best isolated and disposed of via a recycling scheme. Judging by all those YouTube videos, the first signs of battery failure are a burning smell, smouldering fumes or excess heat coming from a device, possibly showing signs of melting plastic. Remove it to the open air immediately and remember that the sound of gas escaping is likely a prelude to a potentially very serious battery explosion which can happen with no warning.

Electrogenic is offering a kit to convert a Land Rover Defender to all electric use. An electric motor is bolted to the Land Rover's existing clutch bell housing, which means the vehicle can keep all its gears, providing up to 90kW and 235Nm of torque. The battery holds 52kWh and is mounted under the bonnet, with a claimed on-road range of 100 miles.

Because of the fire risk, lithium batteries have become a major headache for the airline industry. Hence there are severe restrictions on what batteries and equipment air passengers can 'carry on' and what can be checked into the aircraft hold. For visitors to the US, FAA advice for travellers is at: **https://bit.ly/pe-sep22-haz** while UK travellers and visitors can check the CAA (Civil Aviation Authority) rules at: **https://bit.ly/pe-sep22-caa**

Other news roundup

For some battery applications, Sheffield-based Faradion may eventually offer an alternative to lithium power sources. Their *sodium-ion* batteries are an emerging technology that promises to be safer than lithium cells and offer a higher energy density, and, since they can safely be discharged to 0V, they can be transported more easily with no risk of thermal runaway. Sodium-ion batteries might prove ideal for static storage, back-up power uses or small mobility applications such as e-scooters. There is some outline information at: **www.faradion.co.uk**

Mentioned last month, Hyperloop Transportation Technologies has partnered with Hitachi Rail to develop a digital signalling and traffic management system that will enhance the physical safety requirements for HyperloopTT capsules. It is hoped the tie-up will accelerate the commercial uptake of HyperloopTT by customising Hitachi's proven high-speed rail logic systems rather than creating new standards from scratch.

Owners of Land Rover Defenders could benefit from a new drop-in kit made by Electrogenic that converts these rugged 4x4 diesel workhorses to electric power. The conversion kit, aimed mainly at the agricultural sector and landowners, has been in development and extensively tested over the past 18 months. This very interesting company also specialises in converting old classic cars to EVs. More details at: **https://bit.ly/pe-sep22-lrda**

Rolls-Royce has shortlisted a number of locations around the UK where major parts of its new Small Modular Reactors (SMRs) could be fabricated, including at Grimsby and nearby Stallingborough in north-east Lincolnshire. Other sites in Wales and northern England are also being considered. New headquarters in Manchester have also been announced by Rolls-Royce. It is still early days, with design proposals yet to be signed off before new SMR plants go onstream at the end of the decade.

The SMR race is on: Poland's state-owned energy supplier Enea recently signed an agreement with US SMR developer Last Energy (*Net Work*, June 2022) to co-operate on the deployment of SMRs, while American SMR producer NuScale is gearing up to produce and deliver its 'VOYGR' SMR scalable 77MW modules (*Net Work*, February 2022).

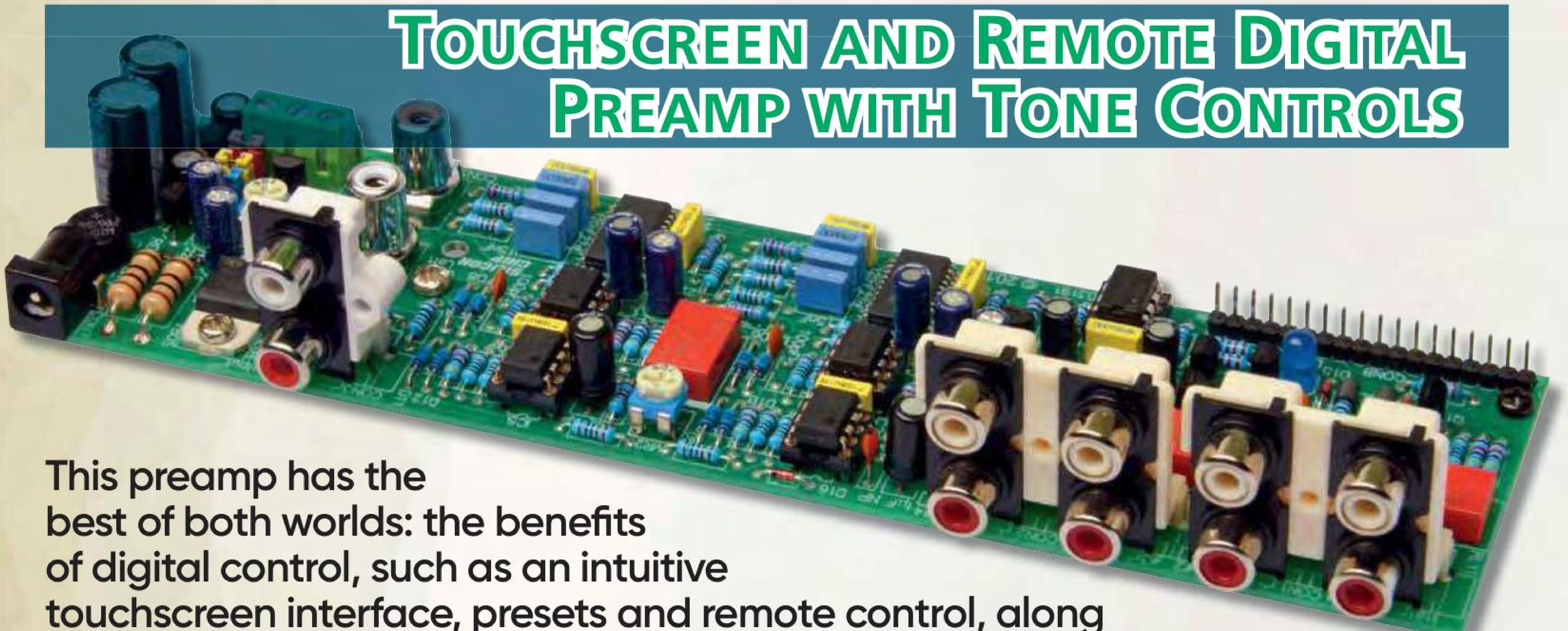
A final reminder that you'll find this month's hyperlinks ready-made for you on the *Net Work* blog page at *PE's* website, at: **www.electronpublishing.com**

That's all from *Net Work* this time, see you next month!

The author can be reached at:
alan@epemag.net

PART 1: BY NICHOLAS VINEN AND TIM BLYTHMAN

TOUCHSCREEN AND REMOTE DIGITAL PREAMP WITH TONE CONTROLS



This preamp has the best of both worlds: the benefits of digital control, such as an intuitive touchscreen interface, presets and remote control, along with the low noise and distortion of an analogue design. It does this by using classic Baxandall-style volume and tone control circuitry with op amps, using high-quality digital potentiometers to provide the adjustments.

To date, most of our remote-controlled preamplifiers have used motorised potentiometers. While these have many benefits, such as low noise and distortion, and the ability to simply turn the knob if you are close to the preamp, they are quite expensive and can be hard to obtain. They can also fail and wear out.

Digital volume control ICs are an attractive alternative, but there have only been a few of these with performance that we would call 'Hi-Fi', and most of those have been discontinued. They also can be pretty expensive and difficult to obtain.

Futhermore, since they only adjust the audio level, we need separate arrangements for input switching (as any self-respecting preamp needs at least a few pairs of inputs) and tone controls. Those are a frequently requested feature for preamps, and we agree that they can be handy. For example, they can compensate for loudspeaker shortcomings, such as a lack of bass or treble, or too much treble. So, any digital preamp we want to desisgn would need to tick the following boxes:

1. Decently low distortion and noise (at least CD quality, and ideally better)
2. Tone controls (ideally with at least three bands for flexibility)
3. A wide volume control range operating in a logarithmic manner
4. Adjustable gain to suit a wide range of signal sources
5. Infrared remote control
6. Input switching
7. Intuitive and attractive colour touchscreen interface for direct control.

We achieved 1) through 4) by using two quad Analog Devices AD8403ARZ10 digital potentiometer ICs. While these are not especially cheap at around £8 each, they are still quite reasonably priced compared to Hi-Fi-quality volume control chips. The eight potentiometers they include let us adjust the volume, bass, mid and treble levels in both channels using just two chips.

These devices have impressive specifications, borne out by our testing, with a rated THD+N figure of 0.003% at 1V RMS/1kHz (they tested considerably better than that), a –3dB bandwidth of 600kHz and an impressively low noise level of 9nV per $\sqrt{\text{Hz}}$. So they are well suited to audio signal processing tasks.

Because each chip has all four potentiometers needed for a channel, the digital pot and its associated op amps are laid out all in one area, simplifying the PCB design and minimising crosstalk between channels.

The input switching is handled by three telecom-style relays, which has worked well for us in the past, as these mechanical devices have minimal impact on signal quality.

Finally, the control interface is handled by a *Micromite LCD BackPack* with either a 2.8-inch, 320x240 or 3.5-inch, 480x320 colour touchscreen. This provides many benefits, such as a nice clear volume readout when you adjust it via the remote, the ability to show the actual frequency response for any given tone control setting and loading/saving presets – the whole nine yards.

It's just what you want for a modern preamplifier or amplifier, without compromising the sound quality.

Besides the *BackPack*, which would generally mount on the unit's front panel (along with the IR receiver), all this circuitry is packed onto a modestly-sized PCB at 206 x 53mm. It has four pairs of onboard RCA inputs, so that it can be mounted at the back of the unit.

It can be powered from a separate AC or split DC supply or an internal transformer with suitable windings. That includes transformers with high-voltage windings to power amplifier modules, and low-voltage secondaries for preamps like this one.

For standalone use, the power input can be an onboard socket on the back, near the inputs, along with the optional rear panel pre-outs. These are in parallel with a pair of internal RCA sockets, which can feed the preamp's output signals to a couple of internal amplifier modules, making a complete preamp/amplifier combination.

Performance

The performance of the preamp is summarised in Figs.1-4. Fig.1 shows a plot of total harmonic distortion plus noise (THD+N) against frequency for an input signal level of 1.5V RMS and an output level of 3V RMS. As the final stage has a gain of two times, this means that the volume control section is set for unity gain.

The 20Hz-22kHz bandwidth plot (in cyan) gives the best indication of audible performance. This shows a total harmonic distortion level of less than 0.001% from around 35Hz up to 2.3kHz. The distortion level rises above 1kHz, with the dashed line showing how the curve would look if the harmonics weren't rolled off at the upper end by the bandpass filter.

As a good CD player is generally expected to have a THD+N figure of less than 0.0018% at 1kHz, we'd say that this preamp exceeds CD quality. That's also indicated by its signal-to-noise ratio of over 100dB, with CDs being limited to 96dB by their 16-bit sampling resolution.

Fig.2 shows how THD+N varies with signal level for some typical gain settings. The rise in distortion at the low end is due to noise being a larger component of the signal for small signals, while the rapid rise at the upper end is where the preamp has run out of headroom and has started clipping.

The best performance is around 2V RMS, a typical level for many playback systems such as CD, DVD and Blu-ray players.

Fig.3 shows how the channel separation varies with frequency. We consider this an excellent result, with worst-case crosstalk of –75dB at 20kHz.

Fig.4 shows the preamp's frequency response with the controls set flat, which only varies by about 0.5dB across the whole audio spectrum, rolling off slightly towards the 20Hz end.

It also shows plots with the bass/mid/treble controls set to their extremes individually. This should give you an idea of the adjustment range that the preamp permits. Of course, you would usually not use the controls at their extremes, as shown in that plot.

Fig.5 shows some more realistic tone control settings (mauve, orange and blue) along with some examples of what happens if you set multiple controls to their maximum extents (red, green and cyan).

Note how there is some interaction between the controls. For example, the treble boost is reduced when a lot of bass or mid boost is introduced. These are somewhat odd situations, though, since you would typically be better off with bass cut instead of using a lot of mid/treble boost, and mid cut instead of a lot of bass/treble boost.

Circuit details

The *Digital Preamp* circuit is shown in Fig.6. Signals are fed into one of four pairs of RCA sockets, CON1A-D and CON2A-D. These have individual 100kΩ termination resistors to prevent signals from deselected devices from floating and causing a thump when switching inputs.

These go to the contacts of a pair of DPDT relays which narrow the signals down to two pairs, and these then go to a third DPDT relay which makes the final selection of which stereo signal reaches the RF filter. The RF filter comprises a 100Ω series resistor, a ferrite bead and a 470pF capacitor to ground for each channel. This RC low-pass filter has a –3dB point of 3.4MHz, while the ferrite bead helps to eliminate much higher frequency signals

Features

- Four input stereo preamp with a colour touchscreen and remote control
- Bass, mid and treble adjustments with presets, plus volume control
- Better than CD quality
- Four external stereo inputs (one active at any time)
- Two stereo outputs, one internal and one external
- Optional loudness control automatically adjusts tone with volume

Specifications

- THD+N: typically less than 0.001%; see Fig.1
- Signal-to-noise ratio: typically around 104dB with respect to 2V RMS input
- Frequency response: 20Hz-20kHz +0,–0.5dB
- Channel separation: >75dB, 20Hz-20kHz
- Signal handling: 0.1-2.5V RMS
- Volume control range: approximately 78dB
- Gain range: −50dB to +27.6dB (0.003 times to 24 times)
- Input impedance: 100kΩ || 470pF
- Bass tone control: ±12.5dB centred around 20Hz (±11.5dB @ 50Hz, ±8.5dB @ 100Hz)
- Midrange tone control: ±11dB centred around 440Hz (±7.5dB @ 200Hz and 1kHz)
- Treble tone control: ±11.5dB centred around about 20kHz (±10.5dB @ 10kHz, ±9dB @ 5kHz)
- Power supply: 12-15V AC, 24-30V AC CT or ±15V DC
- Current draw: typically around 200mA with touchscreen on and <50mA with it off

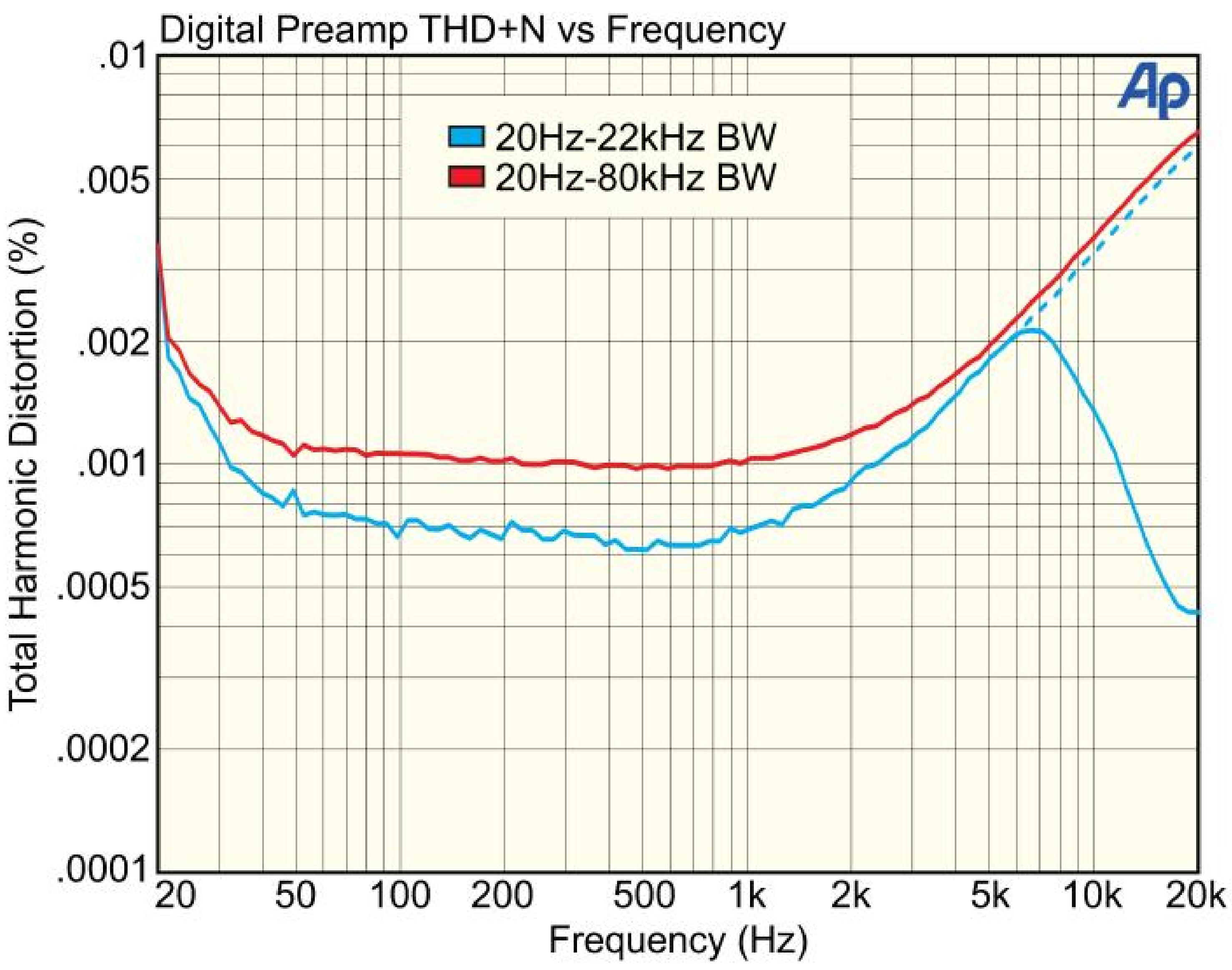


Fig.1: harmonic distortion plus noise plotted against frequency for two different analyser bandwidths. The blue plot with the dashed line is the most realistic representation of the performance, which we think is meritable. 1.5V RMS gives the best performance, but it's still pretty good at around 1V RMS full-scale. The unit can handle over 2.5V RMS at its inputs before clipping.

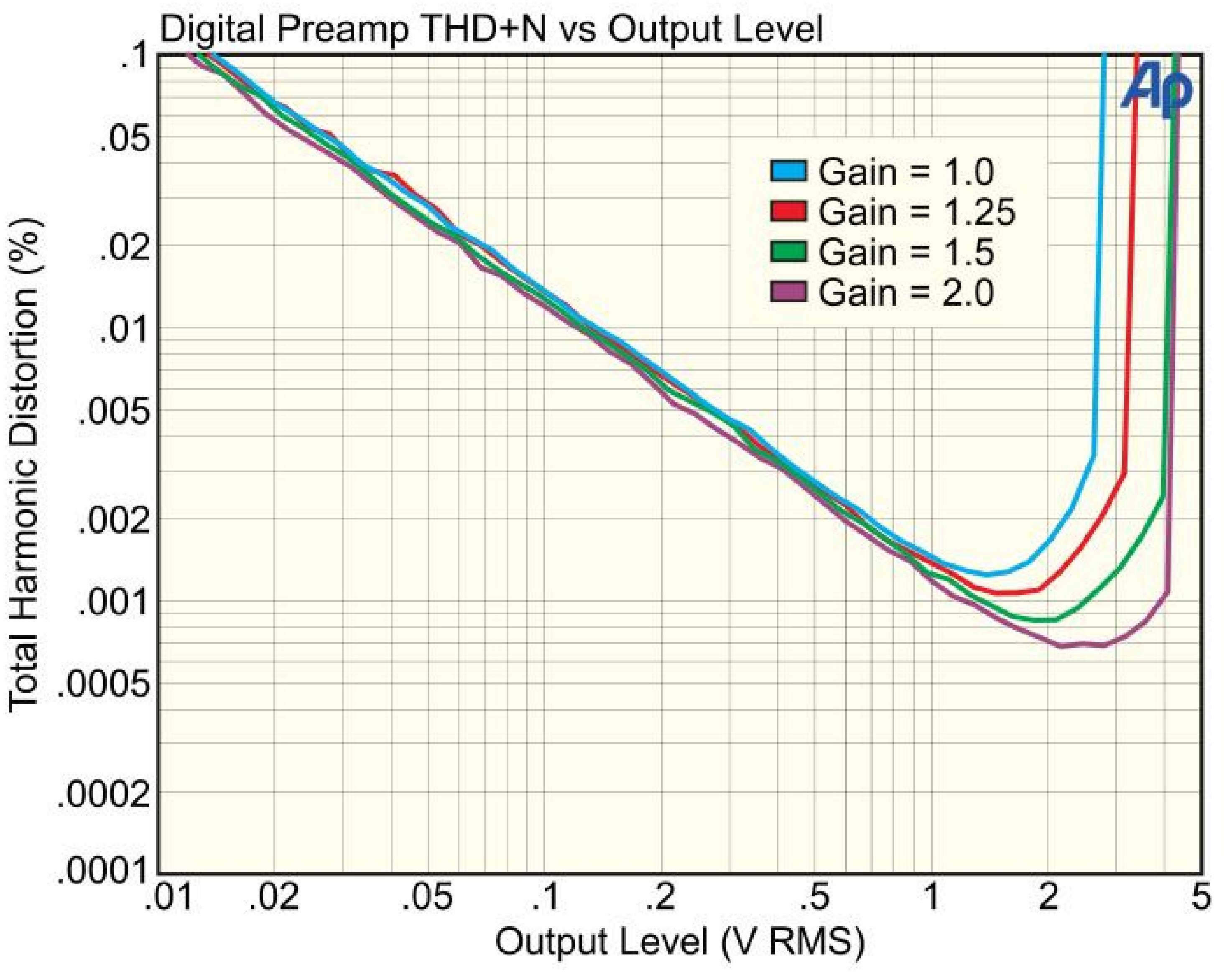


Fig.2: a plot of distortion versus signal level for a 1kHz tone, confirming that distortion rises at lower signal levels due to noise. This also shows the onset of clipping for high signal levels, but note that there are two reasons for clipping; either the input signal rises above 2.5V RMS (as is the case with lower gain settings), or the output runs into clipping at about 4V RMS (higher gain settings).

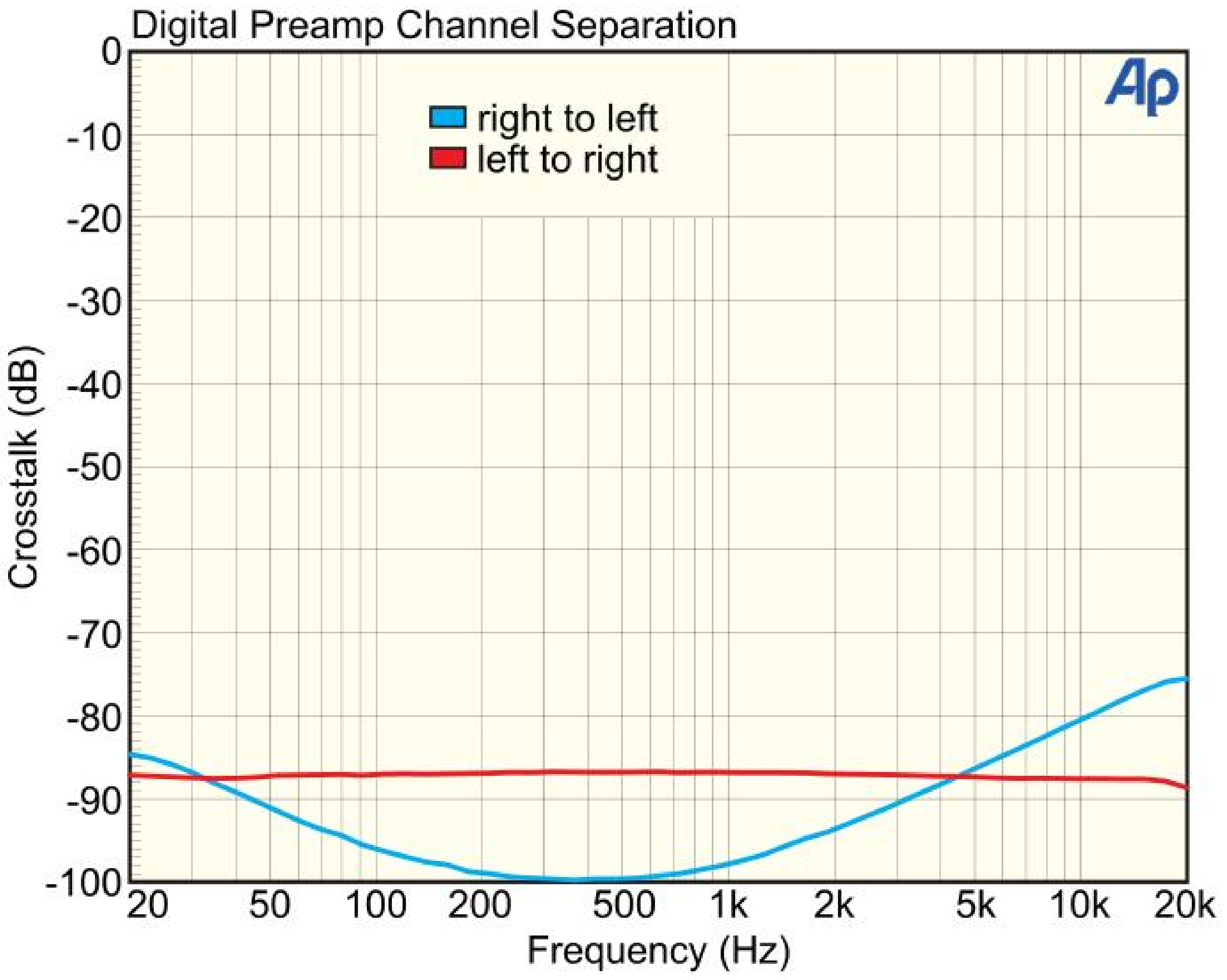


Fig.3: the channel separation of the preamp is excellent, with very little of one channel leaking into the other channel, especially below 5kHz. The input separation is even better, exceeding 100dB in most cases.

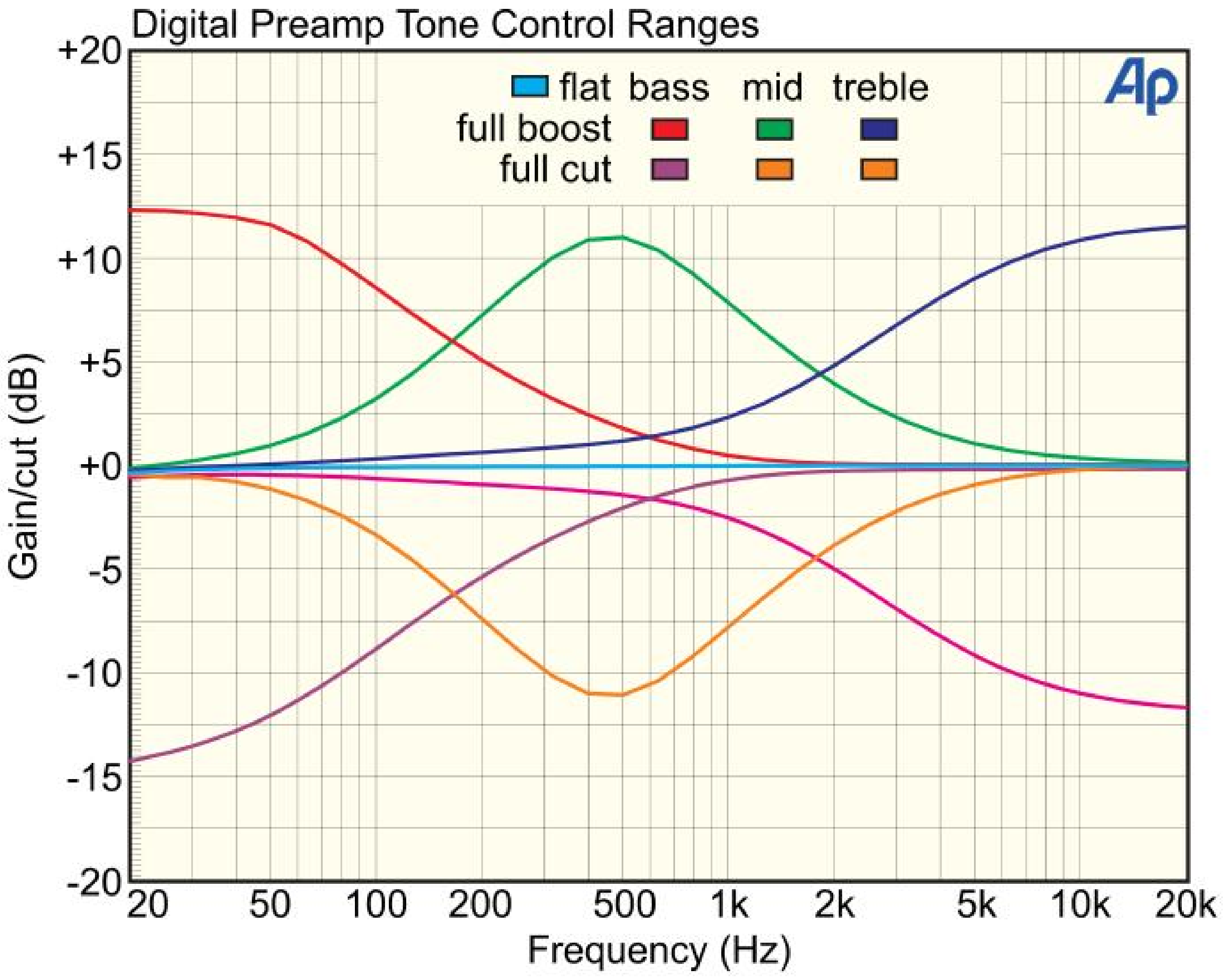


Fig.4: with all the tone settings at 0, the preamp's frequency response is very flat, dropping by only about 0.5dB at 20Hz. The other curves show the result of each tone control being individually set to maximum boost or cut. They indicate how much adjustment you can make and over what frequency range each band operates.

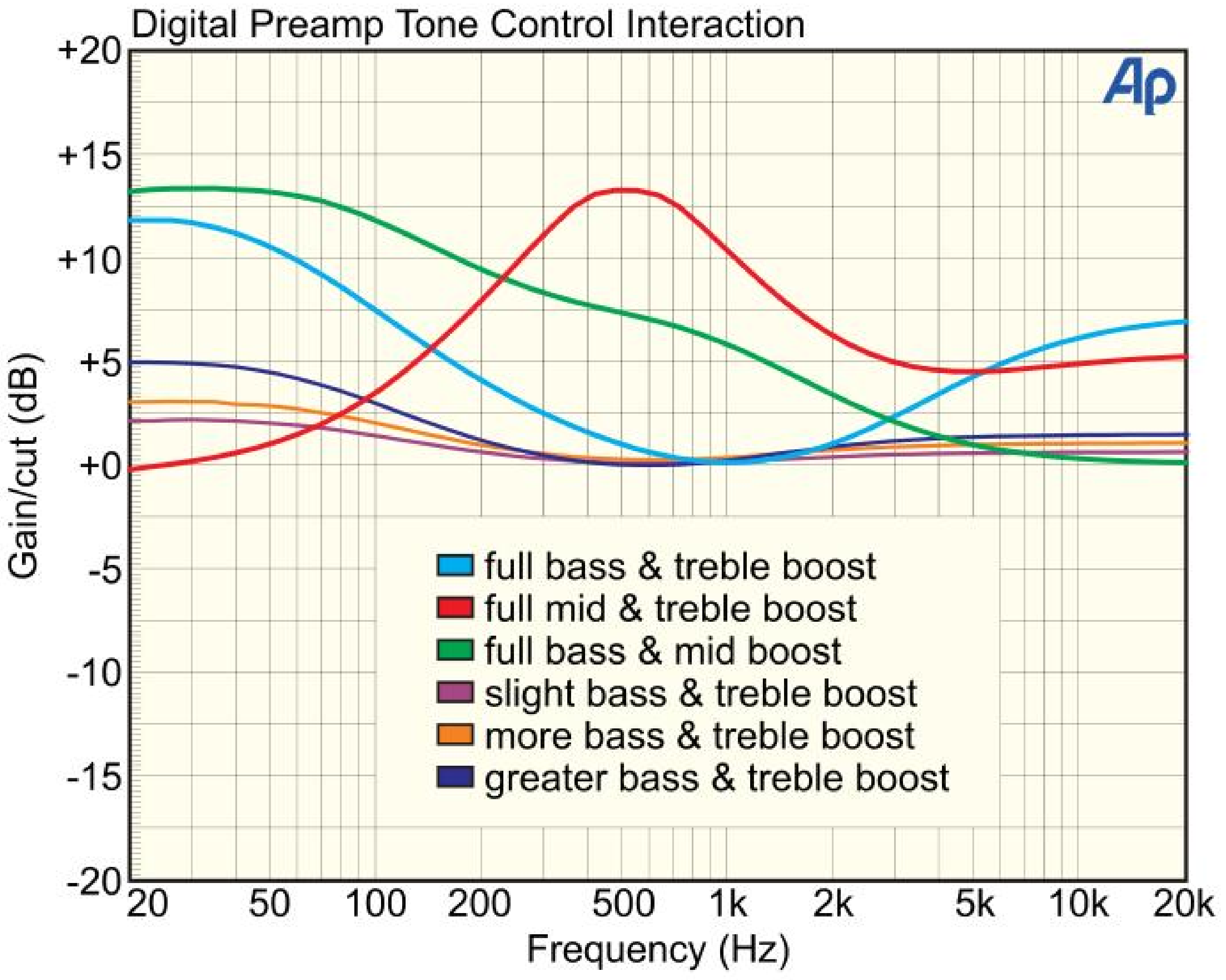


Fig.5: there is a bit of unavoidable interaction between the controls if you make large adjustments in more than one band. The cyan, red and green curves demonstrate this. The other three curves show the results of much subtler simultaneous bass and treble boost settings of various magnitudes. You can see from those curves that there is essentially no interaction at those levels.

which could otherwise be rectified by the following buffer stage, inducing unwanted signals into the audio. 1kΩ stopper resistors further help eliminate RF coupling and also protect op amp IC1 from damage in case a high amplitude signal (or static discharge) is fed into one of the input connectors.

Op amp IC1 buffers the selected stereo signal, and its outputs are AC-coupled to the gain control section via 10µF capacitors. Note that the input side of IC1 is not AC-coupled; it is expected that signals applied to the preamp are reasonably close to 0V DC bias.

The signals from the outputs of IC1 are DC-biased to +2.75V and clamped to be within the range −0.3V to +5.8V. This is done by a pair of schottky small-signal diodes for each channel, connected to ground and a +5.5V rail.

This +5.5V rail is also used to power the quad digital pot ICs, IC6 and IC7. This is their maximum recommended supply voltage (the absolute maximum is +8V). We have done this so it can handle the maximum expected RMS signal voltage from a signal source like a Blu-ray player, which is usually around 2.2-2.3V RMS.

To achieve this, we've had to slightly attenuate the signals being fed to the digital pots, using 2.2kΩ fixed series resistors connected to pin 3 of the two digital pot ICs. These combine with the digital pots' 10kΩ track resistance to reduce the input signals by 18%. So a 2.3V RMS signal is diminished to 1.89V RMS, just within the 1.94V RMS capabilities of the digital pots running from 5.5V.

This is easily compensated for by adding extra gain in the volume control stage. Those 2.2kΩ resistors also limit the current that op amp IC1a needs to deliver if the signal is clipped by diodes D1-D4. IC1a runs from ±12V regulated rails for best performance, so its maximum output swing is about ±10.5V, enough to damage the digital pots without current limiting and clamping.

Volume control

The Baxandall volume control for the left channel consists of dual op amp IC2 plus digital potentiometer #2 within IC6. Similarly, for the right channel, it is op amp IC4 and digital pot #2 in IC7.

Op amps IC2a and IC4a are buffers, while IC2b and IC4b are configured as inverting amplifiers with fixed gains of 14.7 times. The digital pots are then connected within the feedback loop between the output of IC2b/IC4b and the input of IC2a/IC4a. As a result, IC2a/IC4a are fed a signal voltage between that of the input signal and the inverted and amplified output signal.

The net result of this is, with the digital pot 'wiper' (pin 4) all the way at the input (A) end of the 'track', the full input signal is applied to the pair of op amps, so the maximum gain of 14.7 times occurs (actually about 12 times or +21.6dB when you consider the attenuation due to the 2.2kΩ resistors).

As the 'wiper' moves towards the output (B) end of the 'track', the gain reduces logarithmically, eventually to almost zero. The minimum gain (actually attenuation) is limited only by the digital pots' wiper resistances of around 50-100Ω. Our tests show that the lowest gain setting gives about 1.5% of the input signal at the outputs of the volume control section, equivalent to −56dB.

This means that with the volume control at zero, you will still get a little sound out of the preamp, but it will be very quiet. To fully mute the audio, the digital pots have a shutdown feature that disconnects each pot's 'A' terminal entirely. This is where our input signal connects; hence, we can fully mute the outputs if desired.

The output signals from IC2b and IC4b are again clamped to the supply rails by pairs of schottky small-signal diodes, protecting the digital pots from damage if you set the gain too high. The op amps limit the current to around 50mA, so neither the diodes nor the op amp will be damaged during clipping. As the signal is AC-coupled, this will only ever be intermittent anyway.

Tone control

This output signal is AC-coupled to the tone control section via a pair of 47µF capacitors.

The tone control section is the classic Baxandall feedback-based tone control using op amp IC3a for the left channel and IC5a for the right channel. Digital pots #1, #3 and #4 are connected in the negative feedback loops of these op amps, with capacitors connected such that each controls the amount of feedback over a particular range of frequencies.

With these pots all centred, the tone control section has virtually no effect on the signal, basically just acting as an amplifier with a gain of –1. When the pot wipers move off-centre in one direction, signal components in that frequency range are amplified, producing bass, midrange or treble boost. When they move in the opposite direction, signals in those frequency ranges are attenuated (cut) instead.

As the tone control stage is inverting, and the volume control stage is too, the phase of signals fed through the preamp is maintained. Since the outputs of op amps IC3a and IC5a are fed back to the digital pot ICs, they once again are clamped to the supply rails using schottky diodes. The 100pF capacitors directly connecting the outputs to the inverting inputs ensure stability.

Relay RLY4 is the bypass relay. When it is energised, the inverting inputs of op amps IC3a and IC5a are no longer connected to the digital pots. They are instead connected to the centre taps of pairs of 4.7kΩ resistors connecting from the output of the volume control stage to the output of the tone control stage. This configures these two op amps as fixed signal inverters.

The idea behind this is to eliminate any distortion or noise that might be introduced by the digital pots or the associated passive components when a flat response is desired. In practice, the performance of the tone control stage is good enough that this is not necessary.

While we have left provision for RLY4 and its associated components on the board (there would be no real benefit to modifying it to remove them), we don't think the extra cost and complexity is justified. So our parts list and construction details (to come next month) will omit these components.

The output signals from the tone control stages are AC-coupled again, to remove DC bias, then amplified by a factor of times-two by op amps IC3b and IC5b. This allows the output amplitude to be above 1.9V RMS if wanted, up to about 3.8V RMS before clipping. The 100Ω series resistors prevent cable capacitance from affecting those gain stages.

The two outputs are connected in parallel; one is available at the rear panel (if those connectors are

Parts List – Touchscreen Digital Preamp

1 Micromite LCD BackPack programmed with 0110319A.HEX (2.8in display) or 0110319B.HEX (3.5in display) – see below
1 double-sided PCB coded 01103191, 206 x 53mm (from the *PE PCB Service*)
2 double-sided PCBs coded 01103192, 12.5 x 45.5mm (from the *PE PCB Service*)
1 universal IR remote control (**optional**) **[Jaycar XC3718 / Altronics A1012A]**
3 EA2-12 DPDT 12V DC coil telecom relays (RLY1-RLY3)
2 500Ω mini horizontal trimpots (VR1,VR2)
2 small slip-on ferrite beads (FB1, FB2)
3 2-pin headers with shorting blocks (LK1-LK3)
2 quad right-angle RCA socket assemblies (CON1, CON2) **[Altronics P0214]**
1 dual vertical right-angle RCA socket pair (CON3) **[Altronics P0212]**
1 white vertical PCB-mount RCA socket (CON4) **[Altronics P0131]**
1 red vertical PCB-mount RCA socket (CON5) **[Altronics P0132]**
1 3-way mini screw terminal block, 5.08mm pitch (CON6)
1 PCB-mount barrel socket (**optional**) (CON7)
1 18-pin header (CON8)
2 18-pin socket strips
2 16-pin box headers
2 16-pin IDC sockets
1 length of 16-way ribbon cable to suit installation
1 3-pin infrared receiver (IRR1)
1 12-15V AC plugpack/transformer or 24-30V AC centre-tapped transformer with associated wiring, fuse, mains plug etc (to power preamp board)
1 M3 x 6mm machine screw, washer and nut (for mounting REG4)
3 tapped spacers plus 6 machine screws (length to suit installation)

Semiconductors
5 LM833 low-noise dual op amps (IC1-IC5)
2 AD8403ARZ10 quad digital potentiometer chips, SOIC-24 (IC6, IC7) **[Digi-Key, Mouser, RS]**
1 78L12 +12V 100mA linear regulator, TO-92 (REG1)
1 79L12 -12V 100mA linear regulator, TO-92 (REG2)
1 LM317L 100mA adjustable linear regulator, TO-92 (REG3)
1 7805 +5V 1A linear regulator, TO-220 (REG4)
3 PN200 or equivalent PNP transistors (Q1-Q3)
3 PN100 or equivalent NPN transistors (Q5-Q7)
1 through-hole LED (LED1; 3mm or 5mm, any colour)
1 5.6V 1W zener diode (ZD1)
1 W04M bridge rectifier (BR1)
12 BAT42 schottky small-signal diodes (D1-D12)
3 1N4148 silicon small-signal diodes (D13-D15)

Capacitors
2 1000µF 25V electrolytic
3 220µF 16V electrolytic
3 100µF 16V electrolytic
2 47µF 16V electrolytic
2 22µF 16V electrolytic
3 10µF 16V electrolytic
2 1µF 63V MKT
2 220nF 63V MKT
4 150nF 63V MKT
5 100nF 63V MKT
2 33nF 63V MKT
2 470pF ceramic disc
4 100pF C0G/NP0 ceramic disc

Micromite LCD BackPack
We covered the *Micromite BackPack* in *PE* issues May 2017 (V1), May 2018 (V2) and August 2020 (V3). *BackPack* V2 rendered V1 obsolete so the *PE PCB Service* now only stock PCBs for V2 and V3. For all other Micromite requirements we recommend visiting **micromite.org**

Resistors *(all 1% ¼W axial metal film unless otherwise stated)*
11 100kΩ
6 47kΩ
2 22kΩ
2 10kΩ
2 4.7kΩ
6 2.2kΩ
13 1kΩ
1 910Ω
11 680Ω
1 560Ω
1 110Ω
5 100Ω
2 10Ω 1W 5% resistors ***OR***
4 4.7Ω 1W 5% (see text)

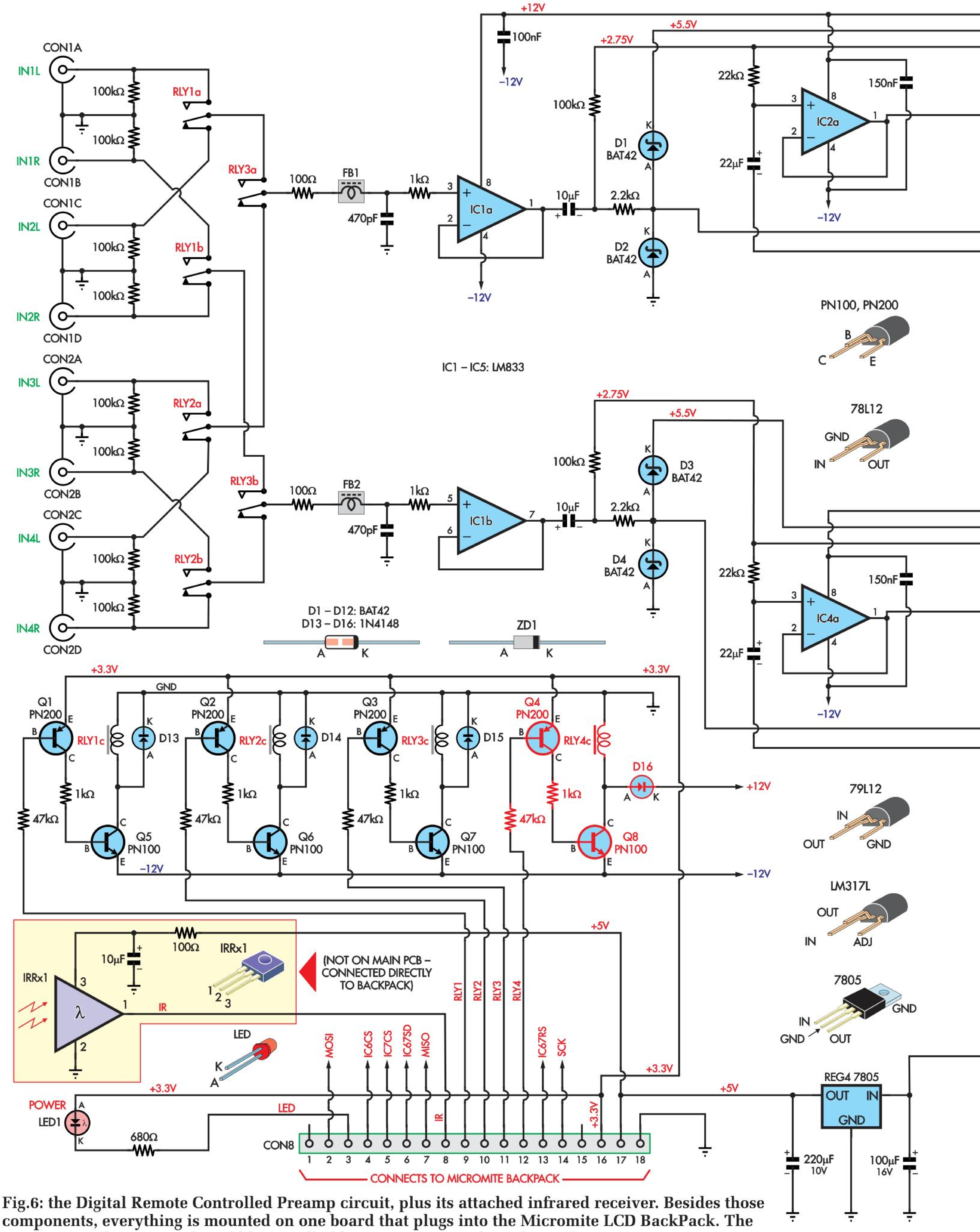
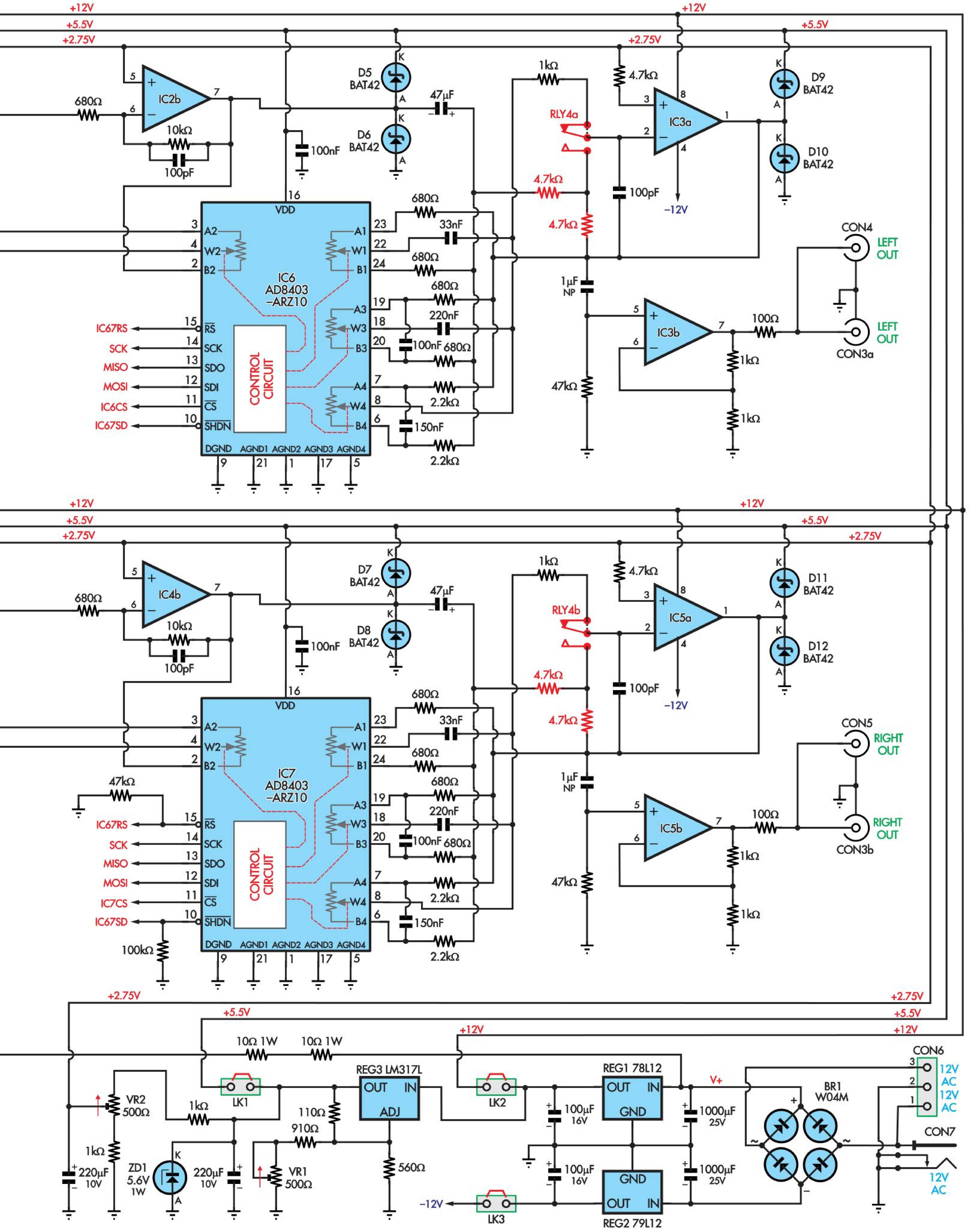

+12V
100nF
–12V
+5.5V
+2.75V
22kΩ
150nF
CON1A
IN1L
100kΩ
RLY1a
100kΩ
IN1R
CON1B
CON1C
IN2L
100kΩ
RLY1b
100kΩ
IN2R
CON1D
RLY3a
100Ω
FB1
1kΩ
470pF
IC1a
100kΩ
D1
BAT42
10µF
2.2kΩ
D2
BAT42
22µF
IC2a
–12V
–12V
PN100, PN200
B
C
E
IC1 – IC5: LM833
CON2A
IN3L
100kΩ
RLY2a
100kΩ
IN3R
CON2B
CON2C
IN4L
100kΩ
RLY2b
100kΩ
IN4R
CON2D
RLY3b
100Ω
FB2
1kΩ
470pF
IC1b
+2.75V
+5.5V
100kΩ
D3
BAT42
10µF
2.2kΩ
D4
BAT42
22kΩ
150nF
22µF
IC4a
–12V
78L12
GND
IN
OUT
D1 – D12: BAT42
D13 – D16: 1N4148
A
K
ZD1
A
K
+3.3V
GND
+3.3V
Q1
PN200
RLY1c
D13
Q2
PN200
RLY2c
D14
Q3
PN200
RLY3c
D15
Q4
PN200
RLY4c
D16
1kΩ
47kΩ
Q5
PN100
Q6
PN100
Q7
PN100
Q8
PN100
+12V
–12V
79L12
IN
OUT
GND
LM317L
OUT
IN
ADJ
7805
IN
GND
OUT
GND
IRRx1
10µF
100Ω
IRRx1
IR
(NOT ON MAIN PCB – CONNECTED DIRECTLY TO BACKPACK)
+5V
RLY1
RLY2
RLY3
RLY4
LED
K
A
MOSI
IC6CS
IC7CS
IC67SD
MISO
IC67RS
SCK
+3.3V
+3.3V
+5V
REG4 7805
OUT
IN
GND
POWER
LED1
680Ω
LED
IR
CON8
1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16 17 18
CONNECTS TO MICROMITE BACKPACK
220µF 10V
100µF 16V

Fig.6: the Digital Remote Controlled Preamp circuit, plus its attached infrared receiver. Besides those components, everything is mounted on one board that plugs into the Micromite LCD BackPack. The components shown in red could be installed but we recommend you leave them off, as our testing shows that they don't provide any benefits.

installed). The other pair consists of vertical connectors on the board, more suited for internal connections to amplifier modules. It should be possible to use both at once, given that the output impedance is relatively low. This could be the case in an integrated amplifier that provides pre-out signals.

Control by Micromite

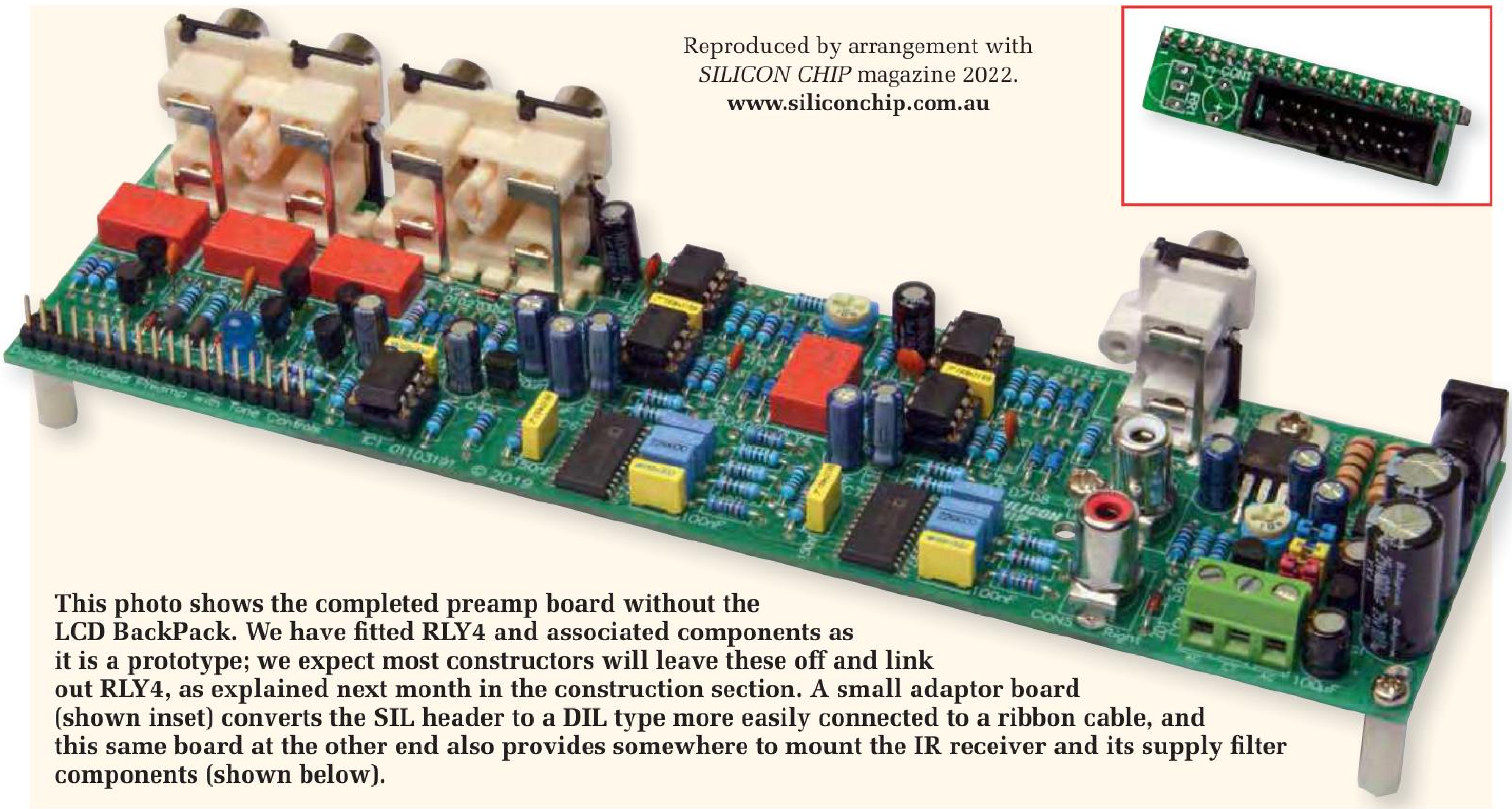
The digital pots are controlled using an SPI serial bus, with one $\overline{CS}$ (chip select) line each, plus active-low


+12V
+5.5V
+2.75V
IC2b
680Ω
10kΩ
100pF
D5 BAT42
D6 BAT42
47μF
100nF
1kΩ
4.7kΩ
RLY4a
4.7kΩ
4.7kΩ
100pF
IC3a
−12V
D9 BAT42
D10 BAT42
IC6 AD8403 −ARZ10
VDD
A2
W2
B2
A1
W1
B1
A3
W3
B3
A4
W4
B4
RS
SCK
SDO
SDI
CS
SHDN
DGND AGND1 AGND2 AGND3 AGND4
CONTROL CIRCUIT
IC67RS
SCK
MISO
MOSI
IC6CS
IC67SD
680Ω
33nF
680Ω
680Ω
220nF
100nF
680Ω
2.2kΩ
150nF
2.2kΩ
1μF NP
47kΩ
IC3b
100Ω
1kΩ
1kΩ
CON4
LEFT OUT
CON3a
LEFT OUT
IC4b
D7 BAT42
D8 BAT42
RLY4b
IC5a
D11 BAT42
D12 BAT42
IC7 AD8403 −ARZ10
IC7CS
47kΩ
100kΩ
IC5b
CON5
RIGHT OUT
CON3b
RIGHT OUT
10Ω 1W
10Ω 1W
LK1
LK2
LK3
REG3 LM317L
OUT IN ADJ
110Ω
910Ω
560Ω
VR1 500Ω
VR2 500Ω
1kΩ
1kΩ
ZD1 5.6V 1W
220μF 10V
220μF 10V
REG1 78L12
REG2 79L12
OUT IN GND
100μF 16V
100μF 16V
1000μF 25V
1000μF 25V
V+
BR1 W04M
CON6
12V AC
12V AC
CON7
12V AC
−12V

Micromite-controlled Digital Preamp with Tone Controls

common reset ($\overline{\text{RS}}$) and shutdown ($\overline{\text{SHDN}}$) lines. That's a total of seven digital lines required to control both of the ICs.

We also have four relays to control. An NPN transistor drives each relay coil with a back-EMF clamping diode. These relays have 12V DC coils, and somewhat unusually, are powered from the −12V rail. This is because the +5.5V rail is derived from the +12V rail, so we are driving the relays from



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This photo shows the completed preamp board without the LCD BackPack. We have fitted RLY4 and associated components as it is a prototype; we expect most constructors will leave these off and link out RLY4, as explained next month in the construction section. A small adaptor board (shown inset) converts the SIL header to a DIL type more easily connected to a ribbon cable, and this same board at the other end also provides somewhere to mount the IR receiver and its supply filter components (shown below).

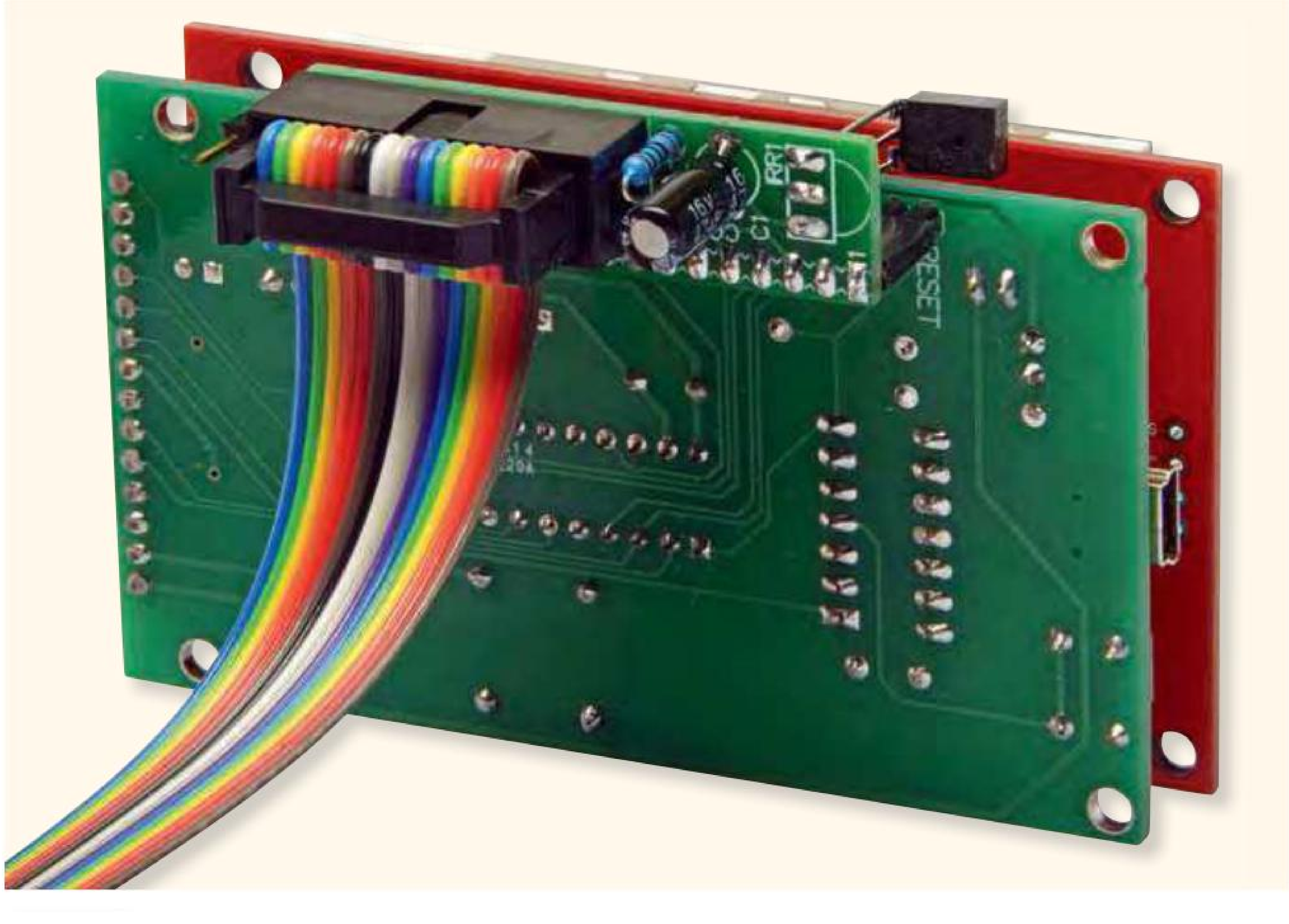
the negative rail to better balance out the current draw. This means that all the relay coil positives are connected to GND, and the negative ends are switched to –12V. Some clamp diodes connect to GND and some to +12V depending on PCB routing convenience; either way, they will still absorb back-EMF spikes and prevent damage to the transistors on switch-off.

PNP transistors Q1-Q4 level shift the 0-3.3V digital relay control signals to allow the NPN transistors with their emitters connected to the –12V rails to be switched normally. So the relays activate when the associated control line is pulled down to 0V, and are off if that control line is at +3.3V or floating (high-impedance).

These 11 total control lines are wired back to SIL header CON8, in positions suitable for being directly wired to the I/O header on a *Micromite LCD BackPack* module.

There are two additional connections: one to allow the *BackPack* to illuminate or flash the onboard LED (LED1) in response to remote control commands and to indicate that power is supplied. This LED could also be duplicated on the front panel, if desired, along with a series current-limiting resistor. The other connection is for infrared reception, at pin 8 of the I/O header.

While the IR receiver and its supply RC filter are shown on the circuit diagram, they are mounted on a small board attached to the *BackPack*, as the receiver needs to be mounted behind a hole on the front panel of the unit.



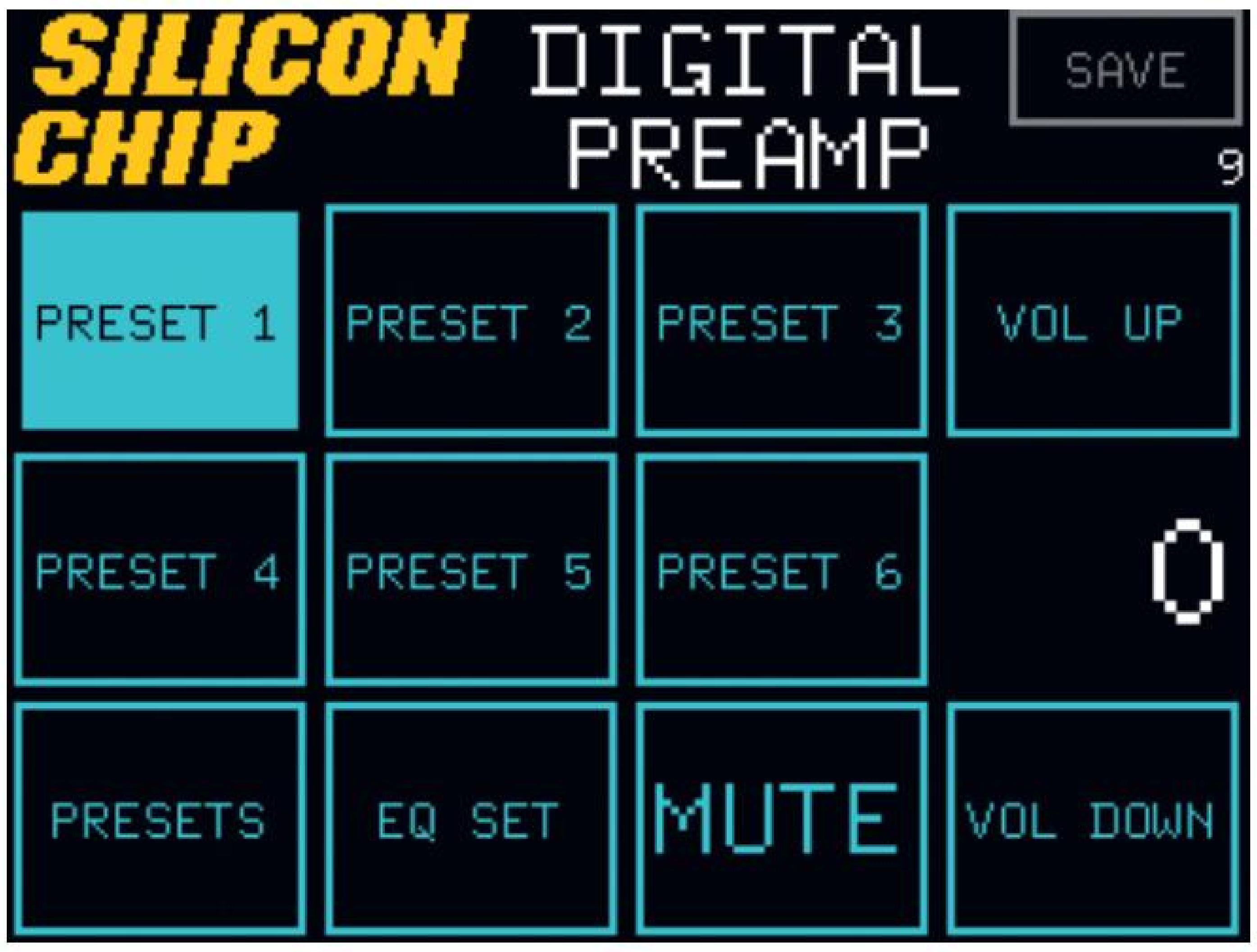
Power supply

The power supply is pretty basic; AC is applied to either barrel socket CON7 or terminal block CON6. If a centre-tapped transformer is used, this would typically be wired to CON6, with the tap to the middle terminal. DC split rails can also be fed to CON6. If AC is applied, this is rectified by bridge rectifier BR1 and filtered by a pair of 1000µF capacitors.

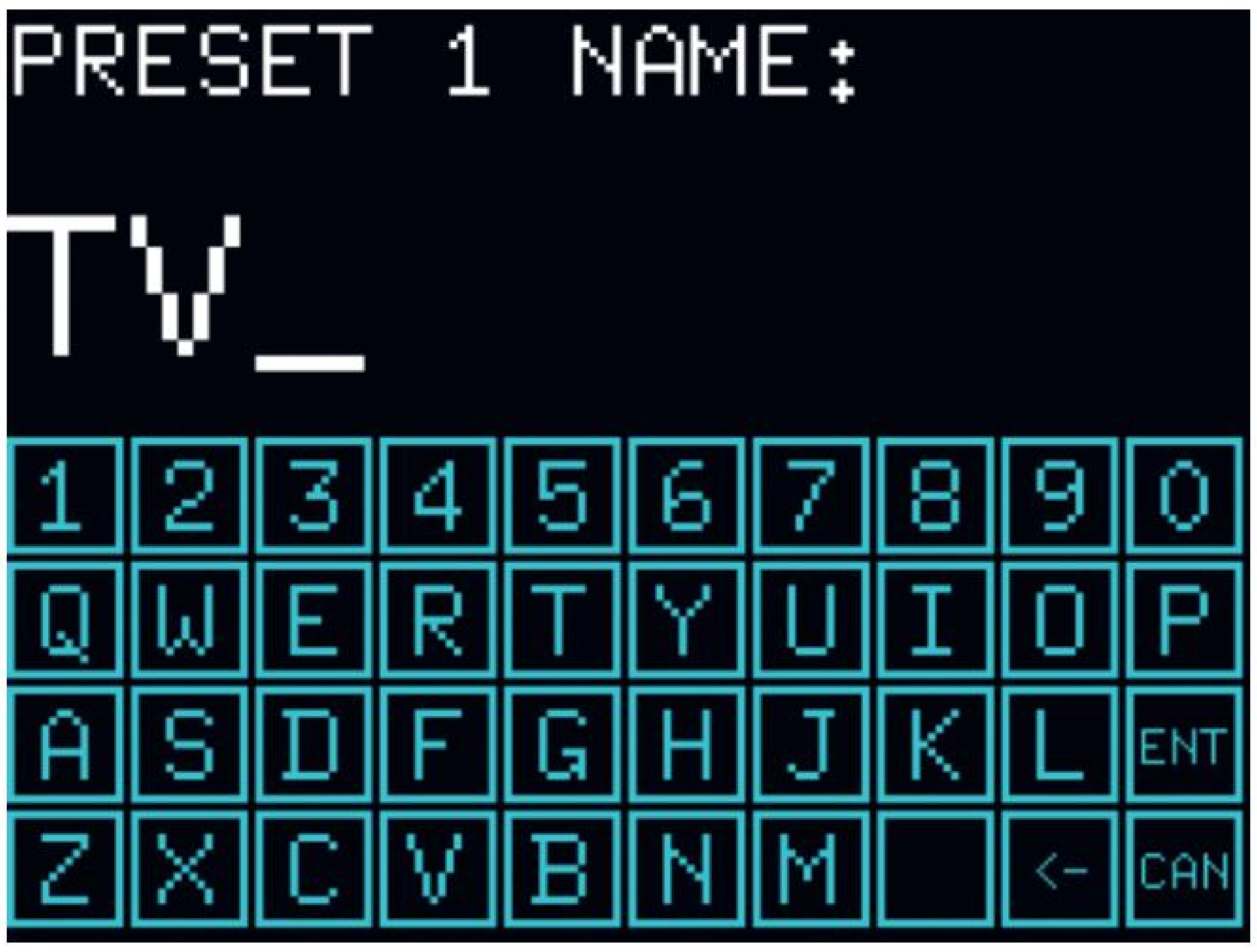
The pulsating DC across these capacitors is then regulated to smooth ±12V DC rails by REG1 and REG2. We have chosen 12V rather than the commonly-seen 15V because the performance is much the same, and we don't need the extra signal swing given the 5.5V limitation of the digital pots. This also provides more headroom for regulation.

The +12V rail is dropped to +5.5V using adjustable regulator REG3. This is adjustable so that it can be set to precisely +5.5V; to be safe, we don't want to exceed the maximum recommended supply voltage for IC6 or IC7 (even though the absolute maximum rating is much higher). A series fixed resistor is provided to limit the adjustment range.

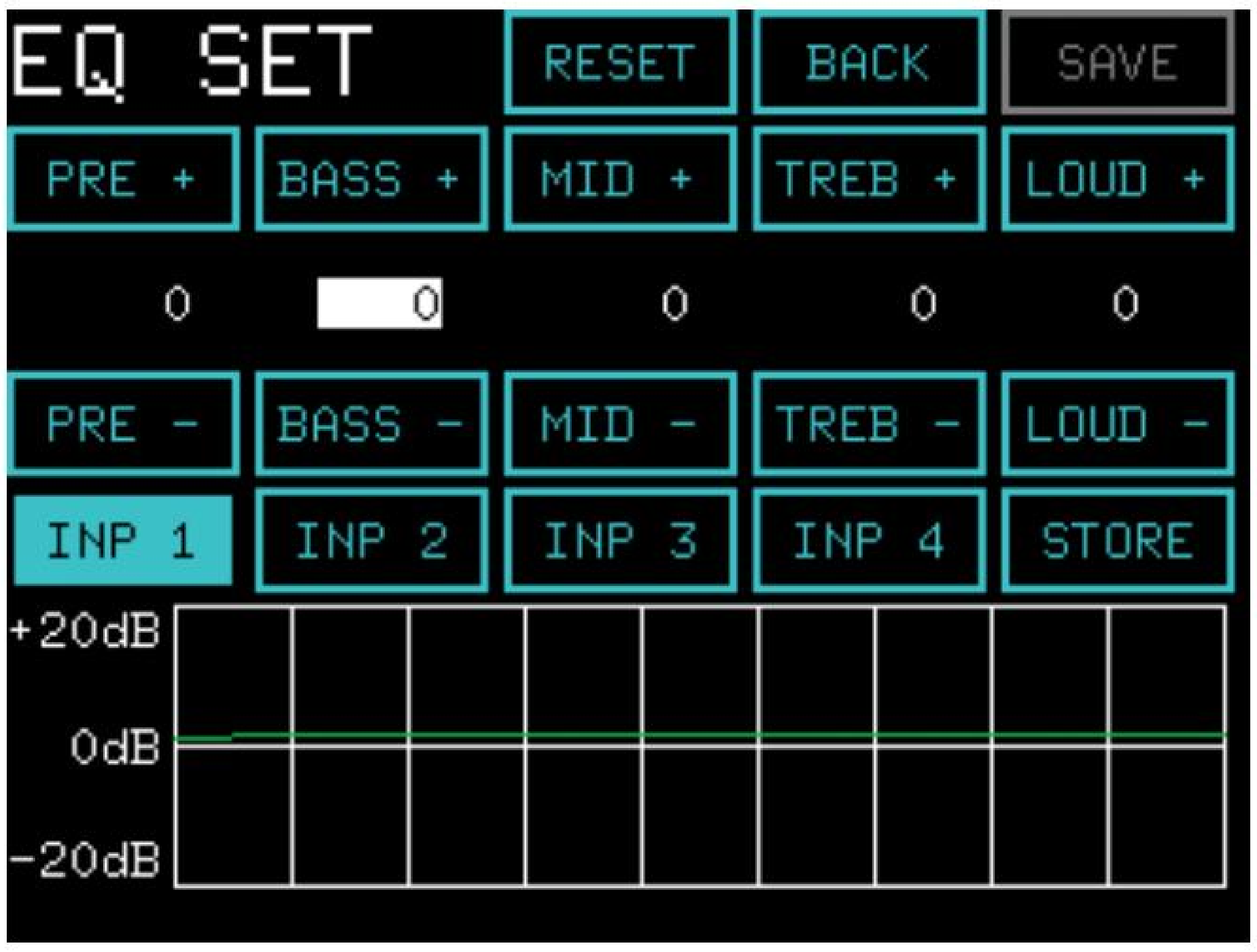
Zener diode ZD1 acts as a safety so that if the output of REG3 is much too high for some reason, it should conduct and prevent damage to IC6 and IC7. The +2.75V mid-supply rail is derived from the +5.5V rail using a resistive divider and trimmed using



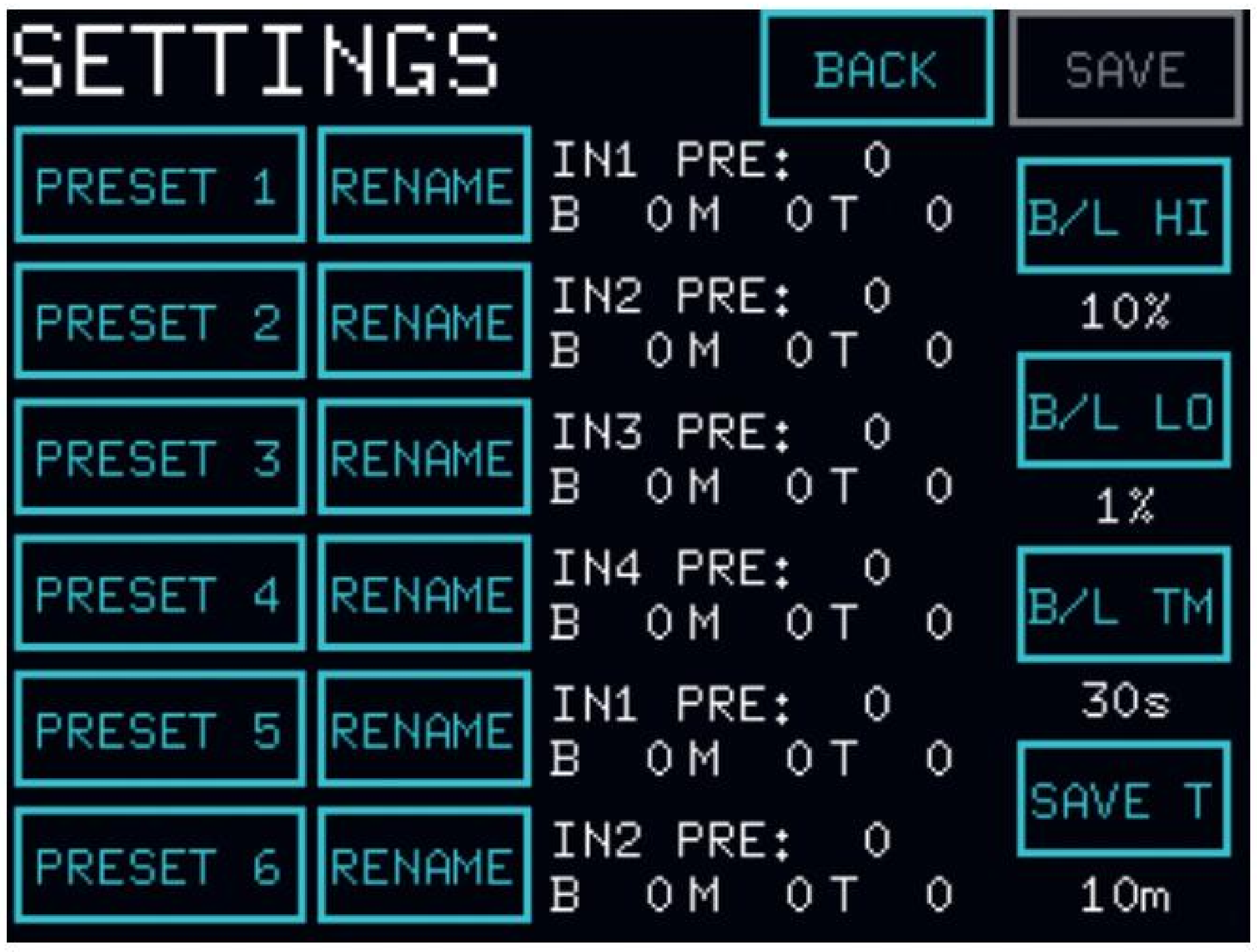
Screen 1: the main screen has buttons to quickly load one of six presets, change the volume, mute the output or go to one of two settings screens (presets and tone/EQ adjustments).


PRESET 1 NAME:
TV_

Screen 4: if you decide to name one of the presets, you will be presented with this basic QWERTY keyboard so you can enter a new name or change the existing one.



Screen 2: the tone control/equaliser (EQ) adjustment screen. Here you can set the bass/mid/treble boost/cut values as well as a volume adjustment (PRE+/–), and it shows you an approximation of the resulting frequency response below. You can also switch between the inputs, adjust the loudness control, reset the settings or store them to the current preset.



IN1 PRE: 0
B 0M 0T 0
IN2 PRE: 0
B 0M 0T 0
IN3 PRE: 0
B 0M 0T 0
IN4 PRE: 0
B 0M 0T 0
IN1 PRE: 0
B 0M 0T 0
IN2 PRE: 0
B 0M 0T 0
10%
1%
30s
10m

Screen 3: in the preset screen, you can switch between the six presets, give them names, view their settings and adjust the backlight brightnesses and timeout.

VR2 so that signal clipping to the supply rails is symmetrical.

It's filtered using a 220µF capacitor so that the source impedance seen by the rest of the circuit is low, preventing unwanted crosstalk.

Links LK1-LK3 are provided for testing because IC6 and IC7 are SMDs. They can be left out while the supply voltages are checked, and IC6, IC7 and the op amps will not receive power. Once the supply voltages have been verified as correct, they can be inserted, and the unit powered back up.

Finally, regulator REG4 provides a 5V DC supply to run the *BackPack* control circuitry. Two series 10Ω 1W resistors have been provided to prevent this regulator from overheating due to the relatively high current required by the *BackPack*, and the large difference in the input (12V) and output (5V) voltages.

This works, although these resistors run fairly hot if you have the *BackPack* LCD backlight turned up to a high brightness setting. If you find this to be a problem, there isn't room to fit a heatsink to REG4, but you could add more dropper resistors. For example, four 4.7Ω 1W resistors mounted vertically instead of horizontally would spread out the heat load.

Software

As the control module is a Micromite, the software is written in BASIC (MMBasic, to be exact). The control program for the *Digital Preamp* is quite small compared to other Micromite-based projects. This is mainly due to the relatively simple functions it provides, with the hardware doing most of the work.

The Micromite processor controls the four relays and the two digital potentiometer ICs, which have four potentiometers each, for a total of eight. The Micromite also commands the LED and receives signals from the infrared receiver.

While the MMBasic code provides an interrupt that is triggered when an IR code is received, we simply use this to set a flag, as other operations could be occurring when the interrupt is triggered. The received command is processed later, when the Micromite would otherwise be idle.

We think that many constructors will want to use the 2.8-inch touchscreen (eg, as used in the original *BackPack* or *BackPack* V2) because it will be a better fit on the front panel of many cases suitable for a preamp.

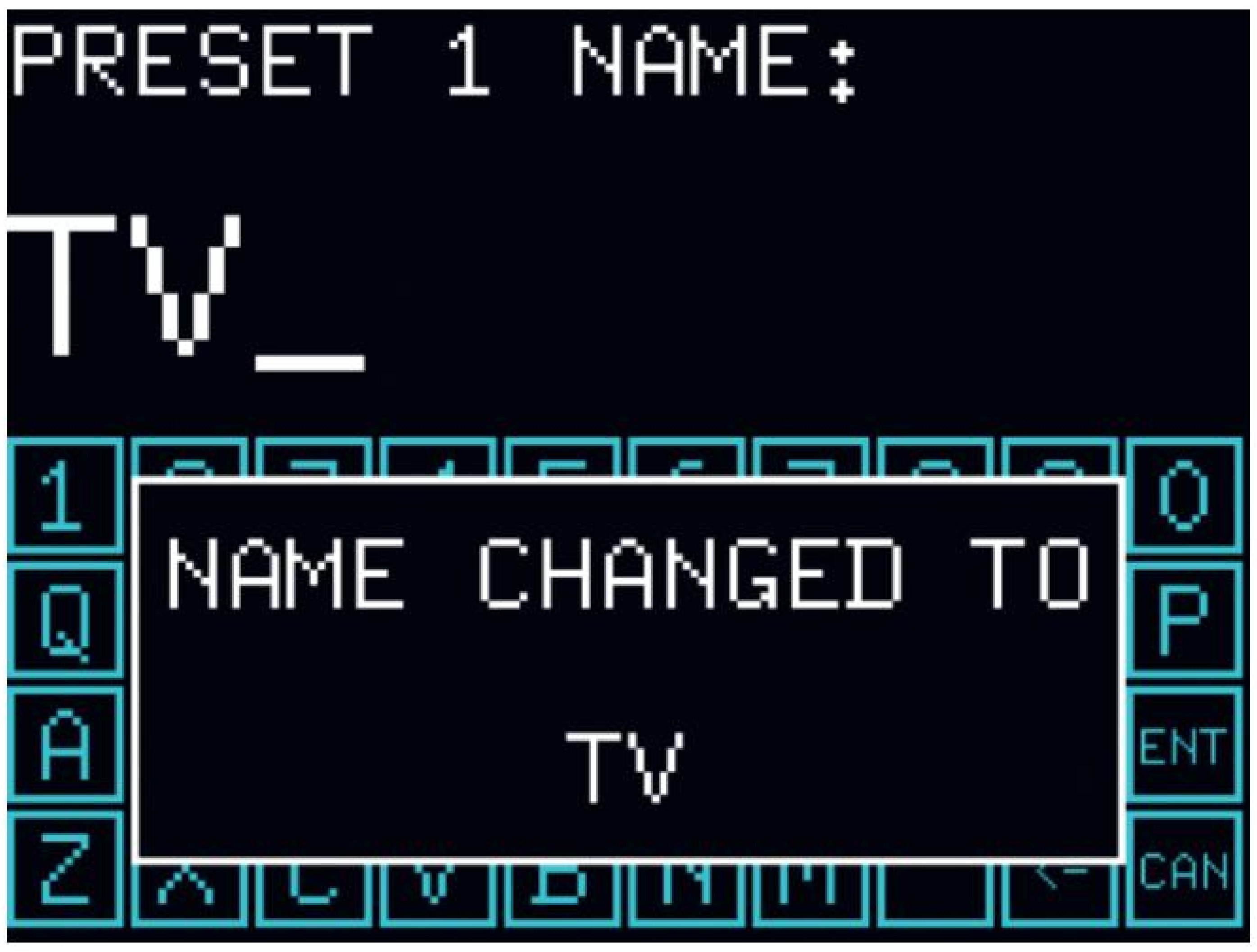
However, you can use the *Micromite LCD BackPack V3* with its higher-resolution, larger 3.5-inch touchscreen if you have room. The software has been designed so that it can use either screen with just minor changes to the code, and we will provide both versions (BASIC code and HEX files) in the download package for this project.

User interface

As with other projects using the *Micromite BackPacks*, several different screens are provided for various features. The MAIN screen offers the features that will be used most often, while two other screens allow the settings to be customised.

The MAIN screen (Screen 1) has six buttons corresponding to six presets. While there are only four inputs, some readers might have these connected to other devices with more inputs, so multiple presets can use the same input to provide various custom tone profiles for each input. You might also want different sound profiles for the same device (eg, to suit movies or music playback).

If one of the presets is selected, its button is highlighted; the MUTE

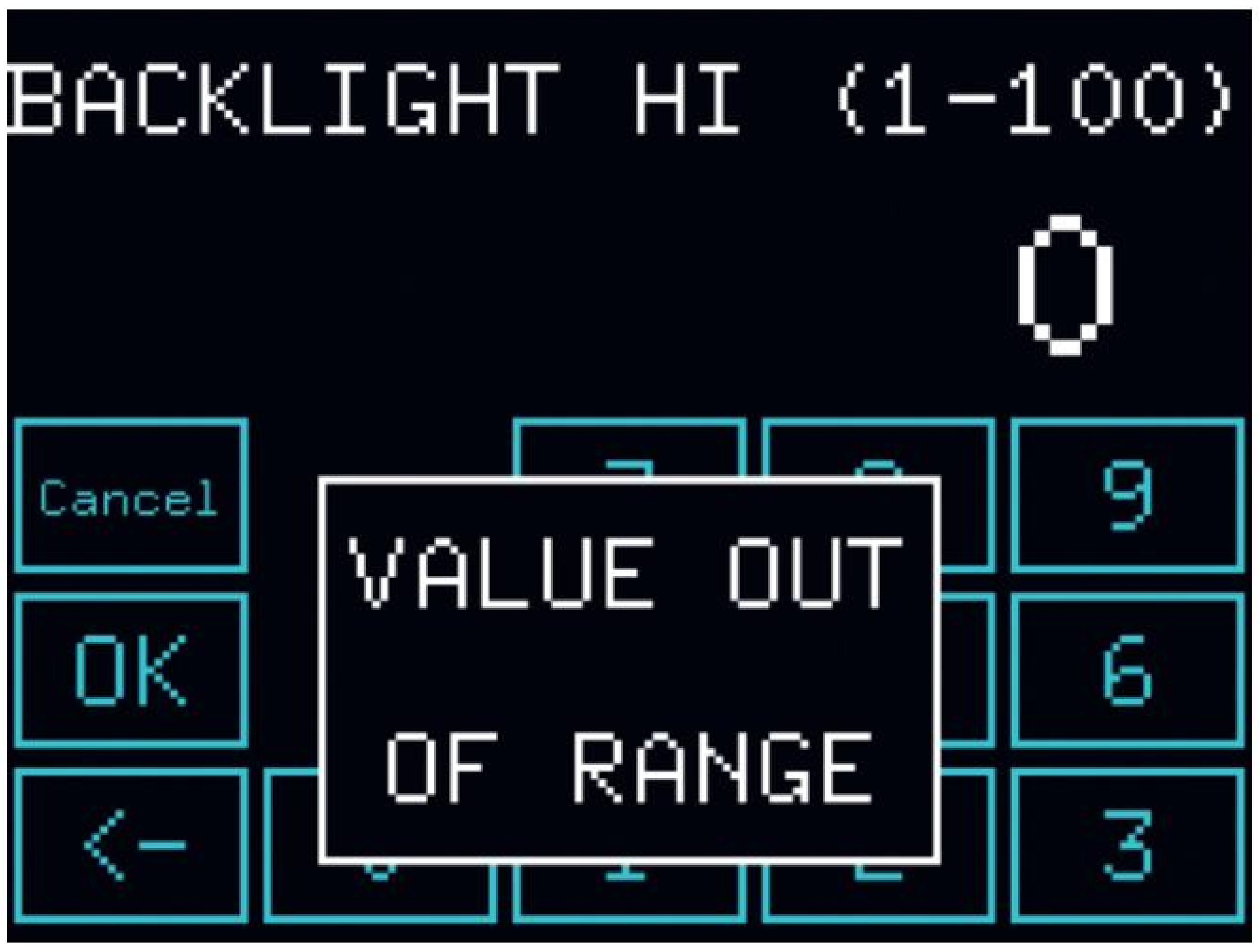

PRESET 1 NAME:

TV_

NAME CHANGED TO

TV

Screen 5: once you press the Enter (Ent) key, it confirms the new name you have typed for the preset.

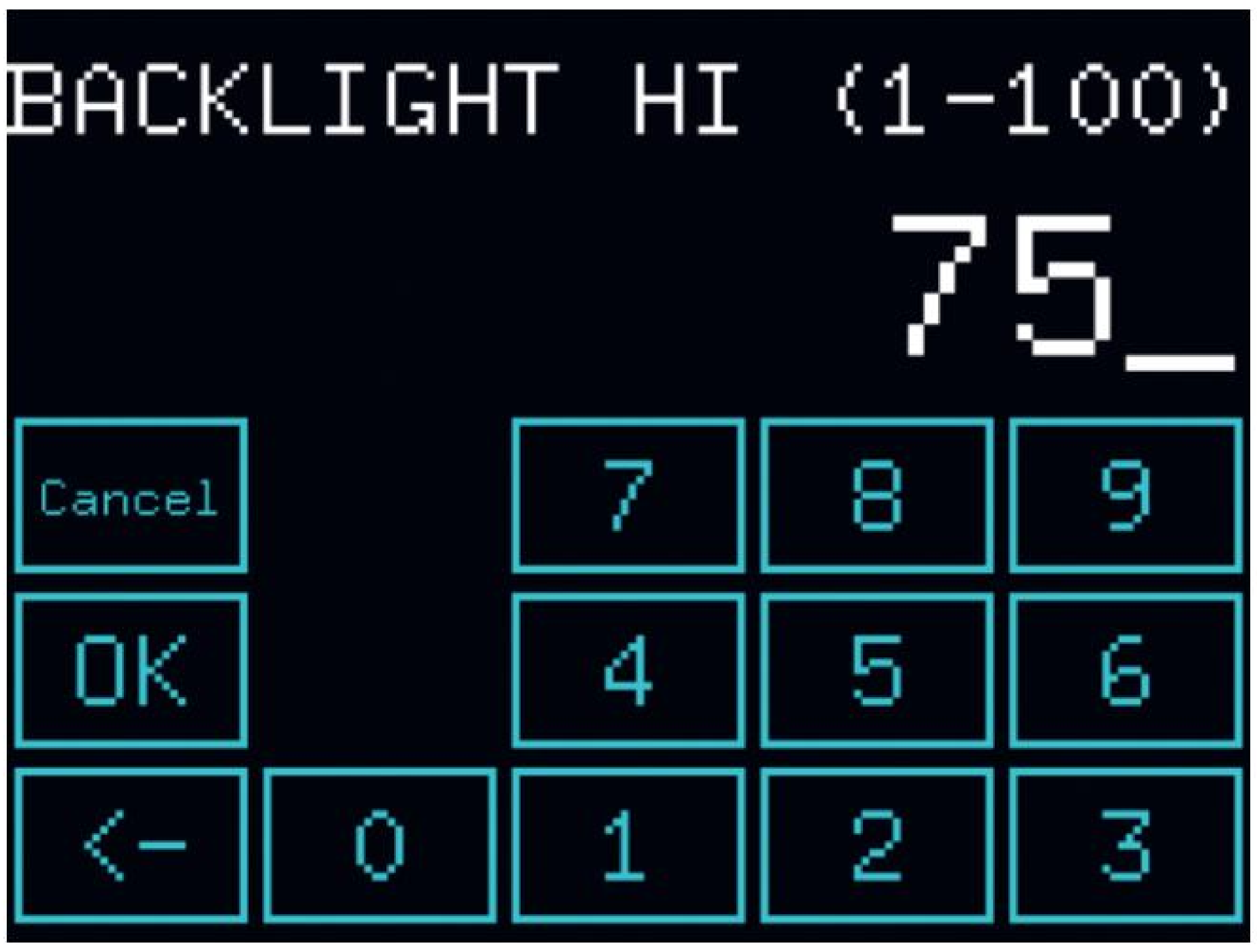

BACKLIGHT HI (1-100)

0

VALUE OUT

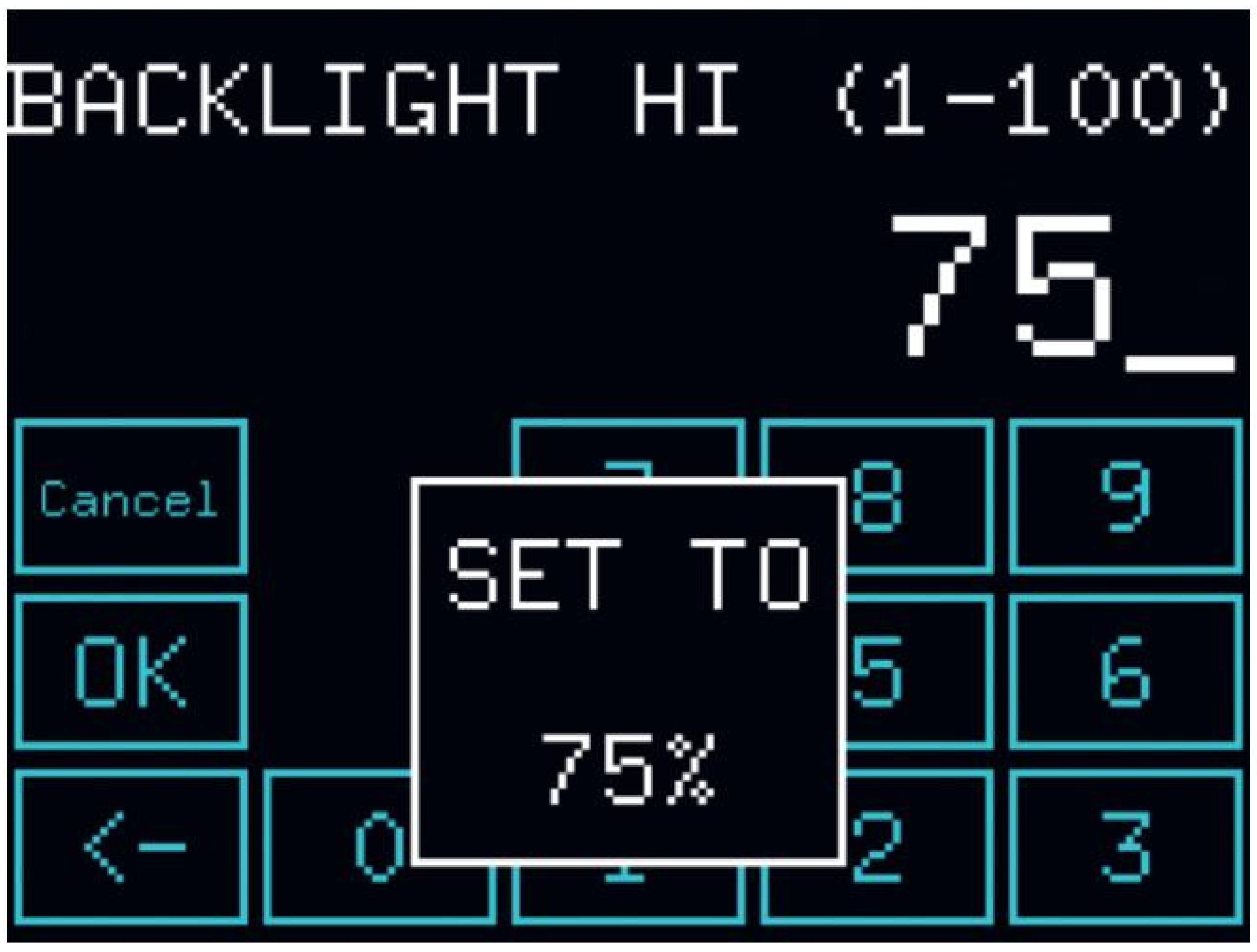
OF RANGE

Screen 8: if you enter an invalid value, an error message will be displayed.


BACKLIGHT HI (1-100)

75_

Screen 6: this simpler numeric keypad is used to enter backlight brightness percentage values. There's one setting for when you are actively using the touchscreen, and another dimmer setting after the timeout. For best audio performance, we suggest using 0% (backlight off) as the timeout value.


BACKLIGHT HI (1-100)

75_

SET TO

75%

Screen 7: the confirmation message that appears when you have adjusted one of the brightness settings.

button is also highlighted when it is active.

Three more buttons provide MUTE, VOLUME UP and VOLUME DOWN functions. These nine buttons correspond one-to-one to the functions that are available via the IR remote control. The level of the volume is also displayed numerically.

At the top right is a much smaller button marked SAVE. Pressing this will cause the current settings to be saved to Flash memory if they have changed. There is also an automatic timed save feature. To help conserve Flash memory longevity, this defaults to 10 minutes (of unsaved changes), but you can alter that.

The SAVE button is red if there are any unsaved changes; otherwise, it is grey. Below the SAVE button is a timer showing the number of seconds before the screen changes to a low-brightness idle mode.

Two more buttons provide access to the settings. The EQ SETTINGS screen (Screen 2) is used to set the tone controls and input selection. This screen also shows an approximate frequency response graph of the current settings.

The graph is based on tests conducted with our prototype, so it will not reflect variances due to component tolerances. The response calculation assumes that the frequency response of each stage is linear, which does not apply at the extreme ends of the potentiometer travel.

The graph is characterised by arrays of values which provide a value for the midpoint response and another value for the difference per potentiometer step at ten different frequencies. These are the values you would need to change if you wanted to characterise your device precisely. The default values should be acceptable for most users.

The SETTINGS (Screen 3) screen allows the currently set tone controls to be allocated to a preset and for these presets to be renamed. The parameters for each preset are displayed next to their buttons. These are shown in raw digital potentiometer steps ranging from −127 to +127, with zero denoting the midpoint.

Each of the six presets can be renamed by pressing the corresponding RENAME button. This brings up a keypad allowing capital letters and numbers to be entered (Screen 4). To make good use of the available space, only a limited set of keys is provided. Backspace, Enter and Cancel buttons are also provided. Upon pressing Enter, the new name is displayed briefly (Screen 5).

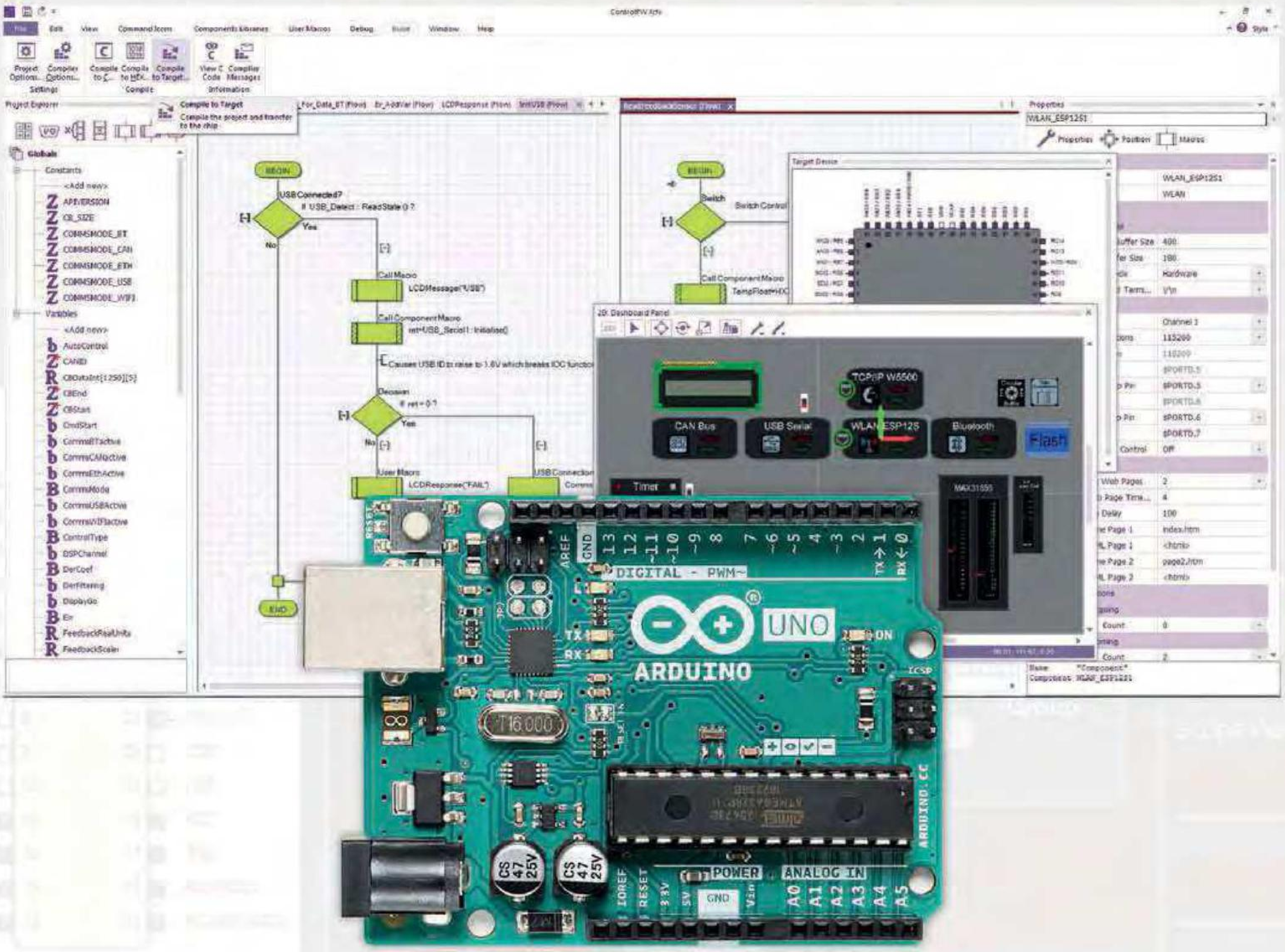
Finally, there are buttons to allow for numeric entry of three backlight settings (normal intensity, idle intensity and idle timeout) and the save timeout setting. Pressing the corresponding button displays a numeric keypad for entering a new value, with the prompt containing a range for valid values (Screen 6).

Entering a value also displays a brief popup indicating the entered value (Screen 7) or noting an error if an entered number is out of range (Screen 8). For simplicity, only positive integer values are supported.

The normal backlight values range from 1-100%, while the idle backlight extends the lower limit to 0%, blanking the display completely. This is handy if you don't wish the display to interfere with, for example, viewing a movie in a dark room.

The idle backlight is only activated on the MAIN screen, so it does not interfere with changing the settings. A touch anywhere on the screen will awaken it; you can use the title area at the top of the screen to be sure of not changing any parameters.

Next month, in Part 2, we will go through the particulars of setup and operation of the *Digital Preamp* after going over the construction and testing details.


Compile to Target
Compile the project and transfer to the chip
BEGIN
USB Connected?
Call Macro
LCDMessage("USB")
Call Component Macro
END
2D: Dashboard Panel
TCP/IP W5500
CAN Bus
USB Serial
WLAN ESP125
Bluetooth
Flash
MAX31855
DIGITAL - PWM~
UNO
ARDUINO
POWER
ANALOG IN

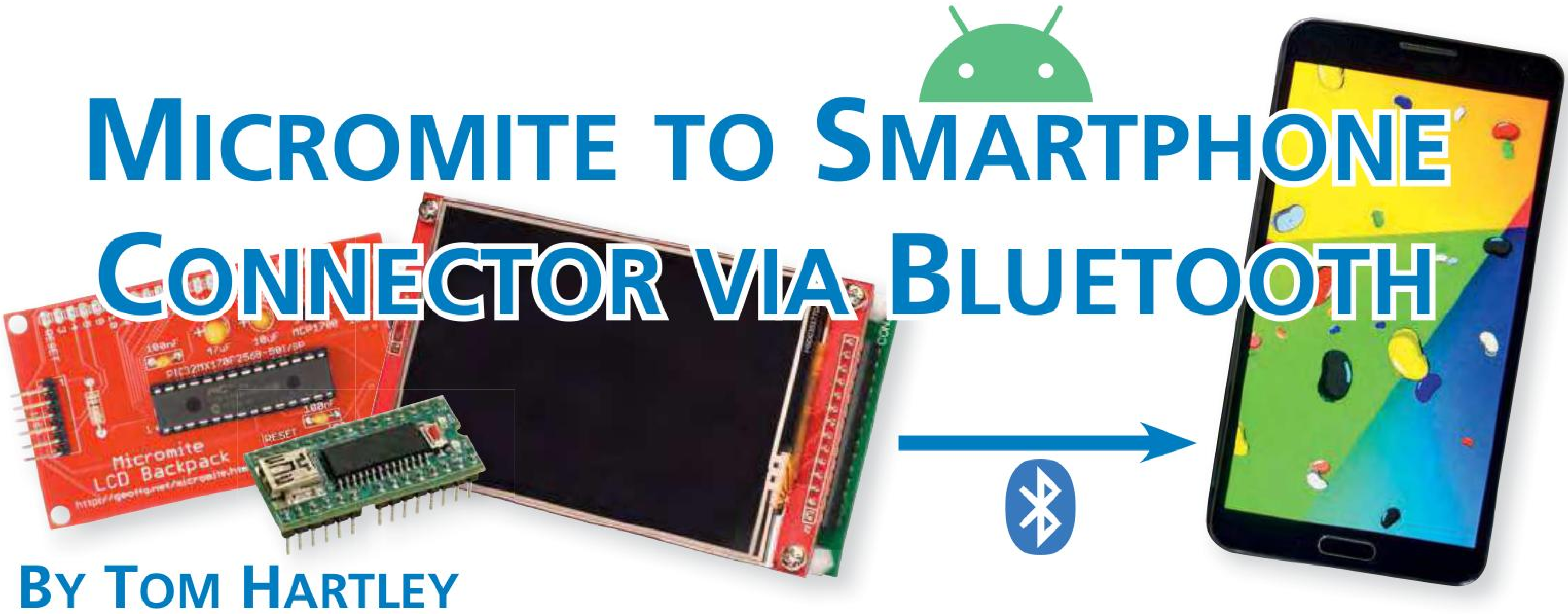
CREATE FANTASTIC ELECTRONIC SYSTEMS USING

FLOWCODE

20% off your Flowcode purchase using code: EPE20

Use code at checkout: flowcode.co.uk/buy

MICROMITE TO SMARTPHONE CONNECTOR VIA BLUETOOTH



BY TOM HARTLEY

This project demonstrates how to use a Micromite as the heart of an IoT (Internet of Things) device. But there are many other reasons you might wish to connect a Micromite to your Android smartphone, such as making it easy to monitor what your device is doing without going to the trouble or expense of fitting it with an LCD screen. It also makes it really easy to control the software running on the Micromite.

Phone Image Source: Android Open Source project

The popular 28-pin Micromite is a great way to get into programming microcontrollers because you need so little to get it up and running, and the BASIC language it uses is easy to learn. But to make the most of it, you really need some sort of screen.

That's why the *Micromite LCD BackPack* series (*PE*, May 2017, May 2018, August 2020) has been so popular. It combines the Micromite with a colour touchscreen, giving you an easy way to interact with the device and display information. But that arrangement is considerably more expensive and complex, and a separate screen isn't always required.

In *Part 14* of *Make it with Micromite* (*PE*, March 2020) we showed how low-cost Bluetooth modules could be used to allow a Micromite to communicate without wires. But what about using such a module to interface with a smartphone?

That way, the phone becomes the user interface to the Micromite, so you can get away with a much simpler and cheaper arrangement – assuming you already have a suitable phone.

And since smartphones generally have a connection to the internet, the Micromite can become an IoT (internet of things) device and easily share data with other devices.

This article explains how to connect a bare Micromite chip to an Android mobile phone to communicate and display data without using a screen. You can even communicate with the Micromite's terminal output data stream using an Android app, sending it BASIC commands and so on.

Basic arrangement

After programming a 28-pin Micromite chip via the conventional PC USB connection, I was able to disconnect it from the PC and transmit the Micromite's terminal output data stream over Bluetooth to an Android App, running on an inexpensive mobile phone. The design requires very few components:

1. A smartphone that is running some version of the Android OS)operating system).
2. A 28-pin Micromite PIC chip loaded with MMBasic, plus a tantalum or ceramic capacitor for the V_{CAP} pin – as recommended by Geoff Graham.
3. An HC-05 Bluetooth module, preferably one with an Enable pushbutton key.
4. A USB-to-TTL converter (eg, one that is based on the ubiquitous CP2102 chip).
5. A short USB extension cable.
6. A BMP180 atmospheric pressure sensor (for this particular demonstration application).
7. A four-AA battery holder modified by tapping the output voltages at 3V and 4.5V. The fourth cell is not needed, so that cell position can be left empty.
8. A small piece of Veroboard.
9. Some hook-up wire.

Bluetooth module setup

The first job is to configure the Bluetooth module as required by this project. The HC-05 Bluetooth module has many similarities to a modem, and the procedure to set it up will be familiar if you have ever set up serial communications to a modem.

Before you can do this, you will need to install a serial terminal

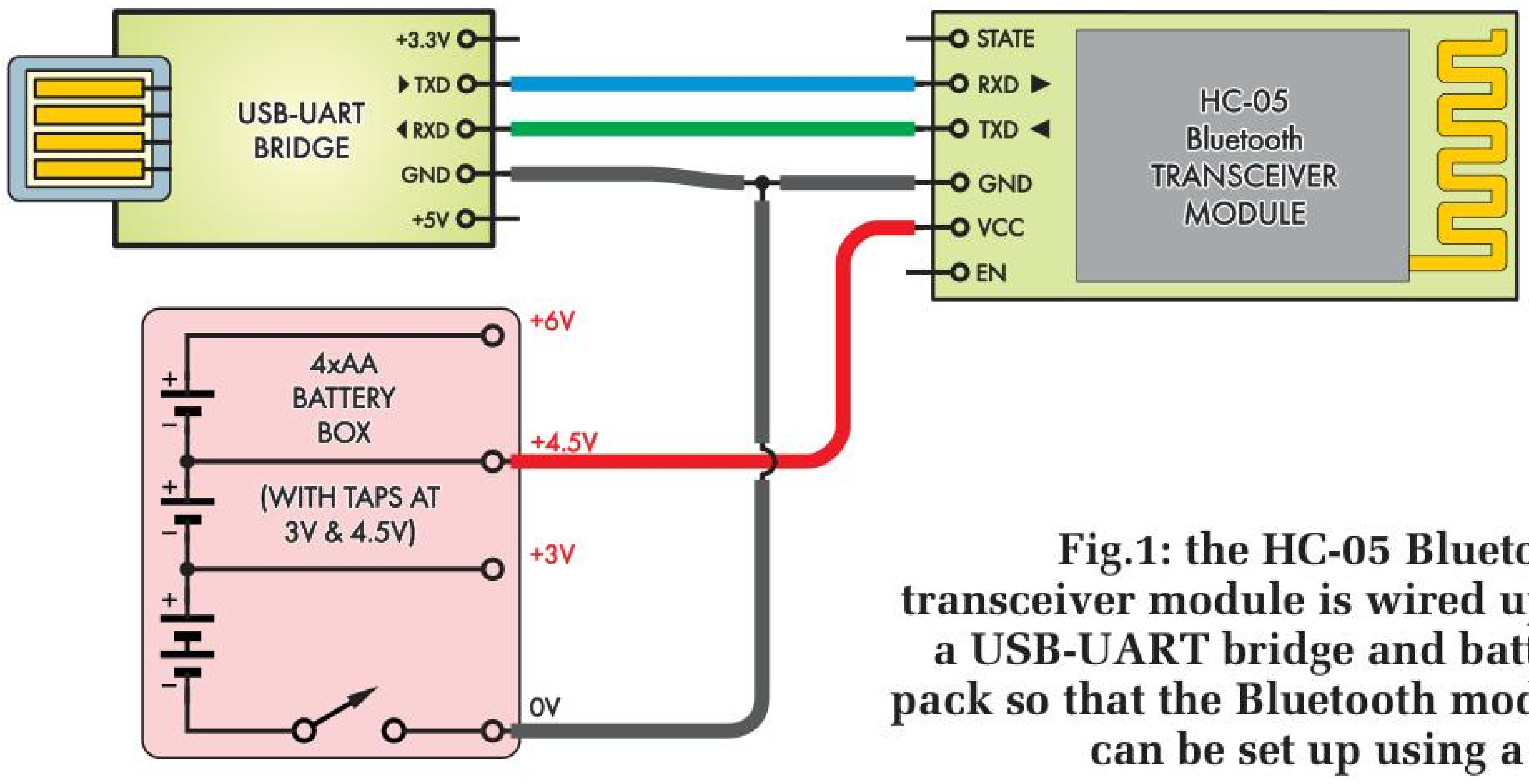


Fig.1: the HC-05 Bluetooth transceiver module is wired up to a USB-UART bridge and battery pack so that the Bluetooth module can be set up using a PC.

program on your computer. For Windows users, Tera Term appears to be the most favoured. For Linux users, the PuTTY SSH Client is recommended. Now, download and install this software.

Next, we need to send the Bluetooth module the appropriate commands to set up the baud rate. These are sent as 'AT' commands. To do this, you have to connect the module to your computer as sown in Fig.2.

Connect the USB-serial adaptor, HC-05 Bluetooth module and battery pack as shown in Fig.2. Start the terminal program on your PC and plug the USB-to-TTL converter into a convenient USB port. This will power up the USB-to-TTL converter but will not power up the HC-05 module.

The terminal software will require information about which USB port it should connect to. You can find this in Windows using the Device Manager. In Linux, when there are no other USB devices plugged into the computer, then the usual USB port is: `/dev/ttyUSB0`

Once you have set that, hold down the button on the HC-05 module and turn the switch on the battery box to the 'ON' position. Wait a couple of seconds before releasing the button. The red LED on the HC-05 module should flash slowly.

Now type `AT` on your computer terminal program and press Enter, the module should respond with `OK`. If it does not, there is probably a baud rate mismatch, so check that the terminal is communicating with the HC-05 at 9600 baud, 8 bits, no parity, one stop bit, no flow control (often described as '8-N-1').

Also, the Enter key on your PC must be mapped as a carriage return plus line feed, usually signified in the terminal software as CR/LF. The other baud rate to try is 38,400. Different manufacturers have different default baud rates on first use. Once you get the OK, you can proceed to enter these two commands:

```
AT+UART=38400,1,0
AT+NAME=MMITE01
```

You should get an `OK` after each one. If you don't, you might have a different version of the HC-05 Bluetooth module; see the panel below.

Next, check that the settings have been recorded by typing '`AT+UART`' and pressing enter, which should provide the response `38400,1,0`. Then type `AT+NAME` and press Enter; you should give the response `MMITE01`.

Power off the circuit and install the HC-05 in the test rig described in the next section.

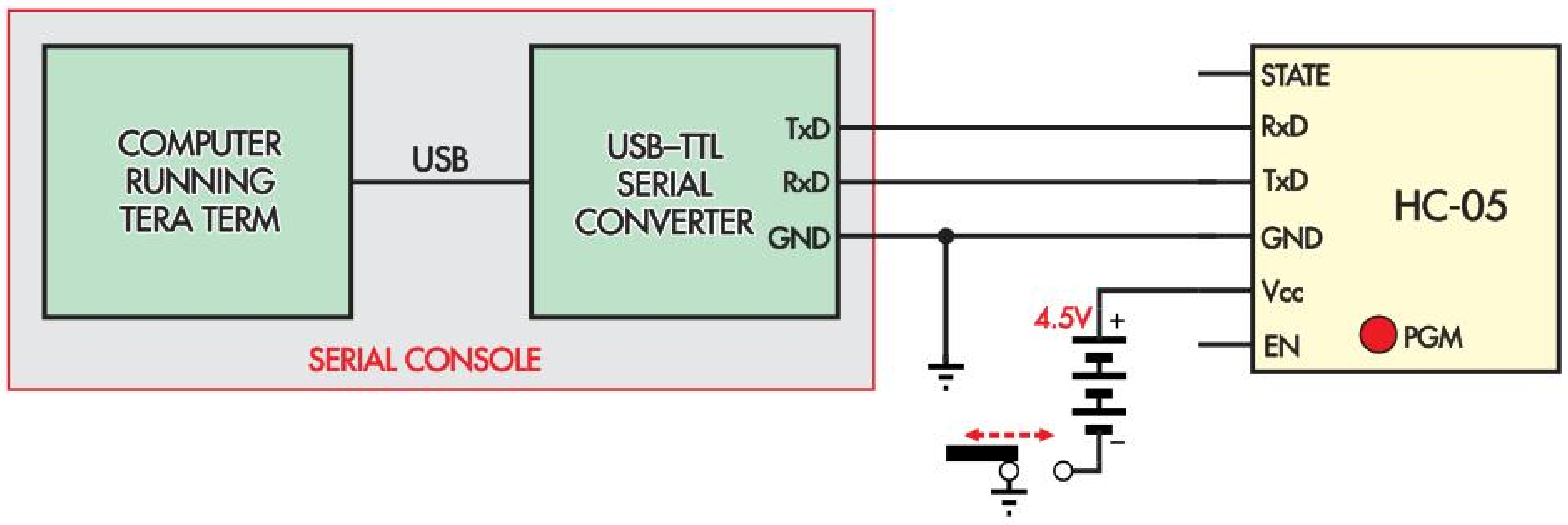


Fig.2: you need to change some settings in the HC-05 Bluetooth module before using it, via serial commands from a computer. This is how you can connect it up in order to do that. The suggested wiring is in Fig.1.

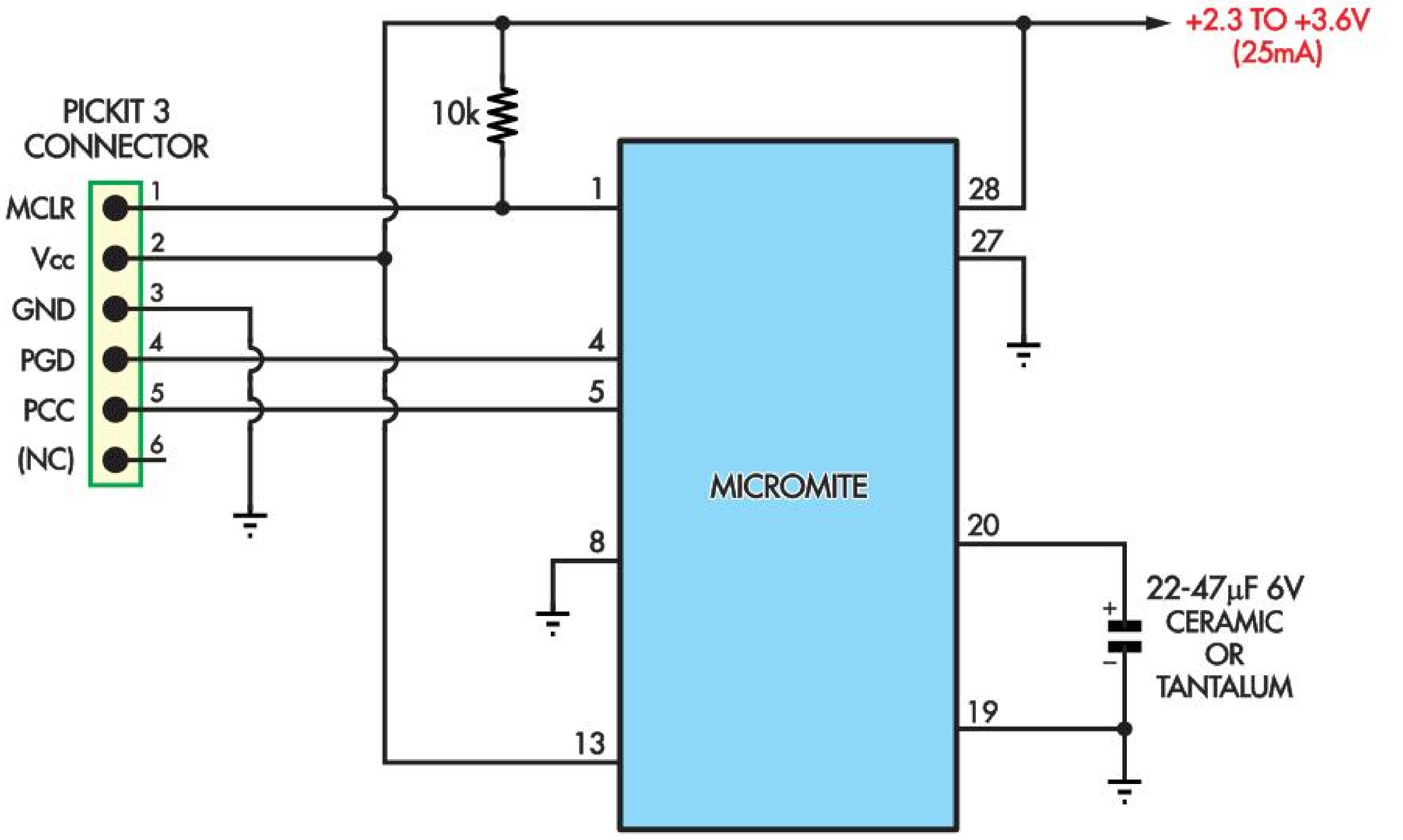


Fig.3: this minimal circuit is all you need to load the MMBasic firmware onto a PIC32, turning it into a Micromite. You can buy an inexpensive pre-programmed 28-pin Micromite from: https://micromite.org/micromites/

Next, install the Bluetooth Terminal app by Kai Morich on the smartphone. You can download it from: **https://bit.ly/pe-sep22-sbt**

Building the circuit

Fig.3 shows how to load the firmware onto the PIC32 chip using a PICkit if it is not already loaded.

Fig.4 is the minimal circuit to build so that you can interface with the Micromite running MMBasic.

However, you won't be able to do much with such a basic configuration, so we will describe how to get the circuit shown in Fig.5 up and running. This includes a BMP180 temperature/atmospheric pressure sensor so it can actually do something useful.

Note that with the Tx/TxD lines of the two serial modules in parallel, you can only have one active at a time. That's assuming that the inactive module is not driving its Tx line actively, which is the case with the HC-05 and USB-serial modules I used, but might not be true for all such devices. If both Tx lines are active at the same time, it's unlikely anything will be damaged (although not impossible), but it certainly isn't going to work as they will fight each other.

Alternative versions of the Bluetooth module

We have seen online sellers listing various versions of the HC-05, including the 'original' version (likely the one described in this article), a 'new' or 'revised' version and the HC-06.

We ordered some of the new/revised HC-05 modules to try out. They look much the same as the original HC-05, and if you order one from a seller who doesn't make the distinction, that may well be the one you receive.

The new/revised version worked as described in this article, except that it did not respond to the 'AT' commands listed in this article at all. However, it seemed to default to 38,400 baud, so we were able to communicate with a Micromite simply by wiring it up and setting that as the baud rate.

We haven't tried the HC-06, but chances are it works much the same way. You might just need to experiment with the baud rate if you cannot communicate with it after selecting 38,400 baud.

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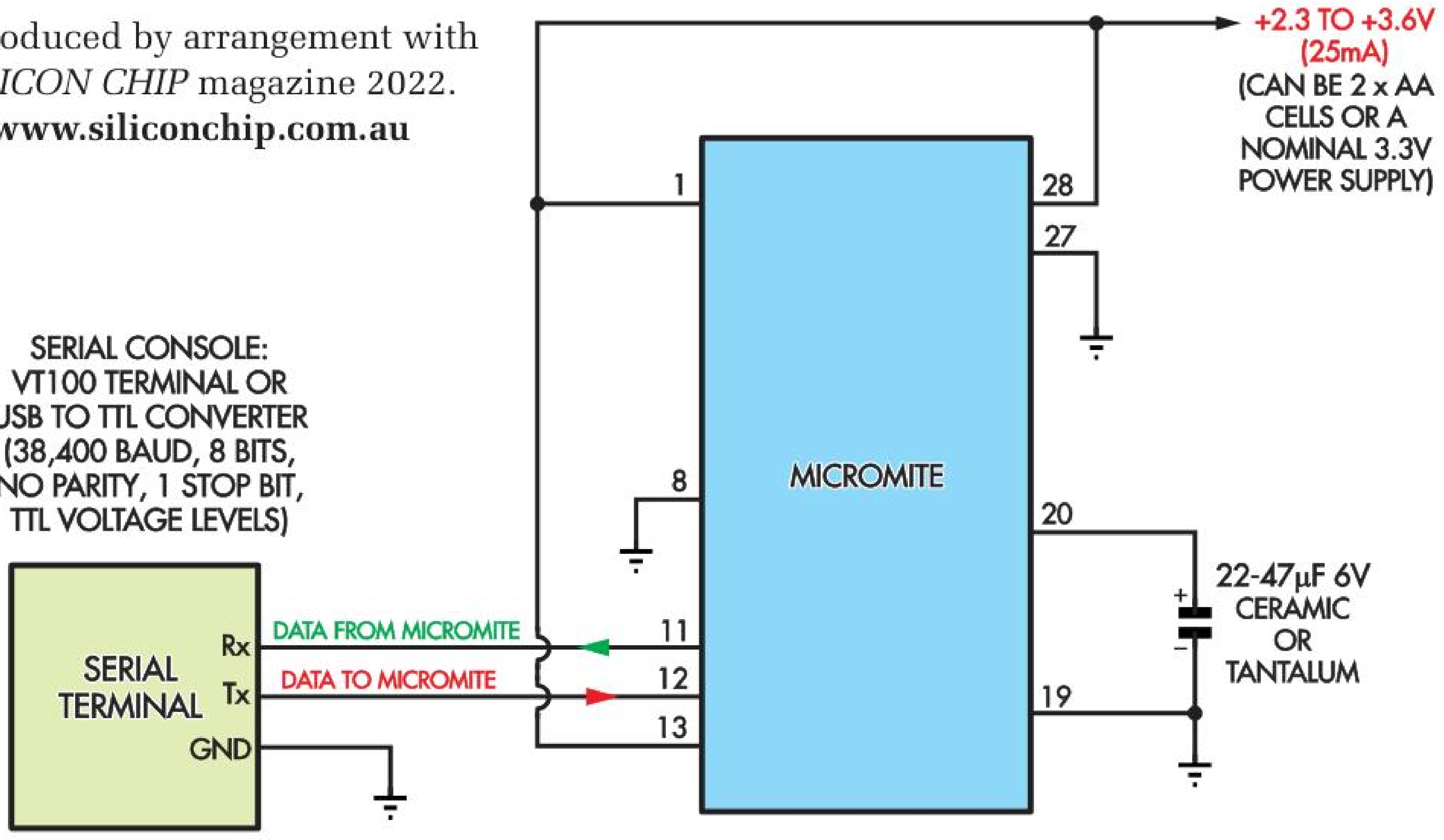

+2.3 TO +3.6V (25mA)
(CAN BE 2 x AA CELLS OR A NOMINAL 3.3V POWER SUPPLY)
1
28
27
SERIAL CONSOLE: VT100 TERMINAL OR USB TO TTL CONVERTER (38,400 BAUD, 8 BITS, NO PARITY, 1 STOP BIT, TTL VOLTAGE LEVELS)
8
MICROMITE
20
22-47µF 6V CERAMIC OR TANTALUM
SERIAL TERMINAL
Rx
Tx
GND
DATA FROM MICROMITE
DATA TO MICROMITE
11
12
13
19

Fig.4: the minimal circuit to communicate with the Micromite over USB, using a USB/Serial adaptor.

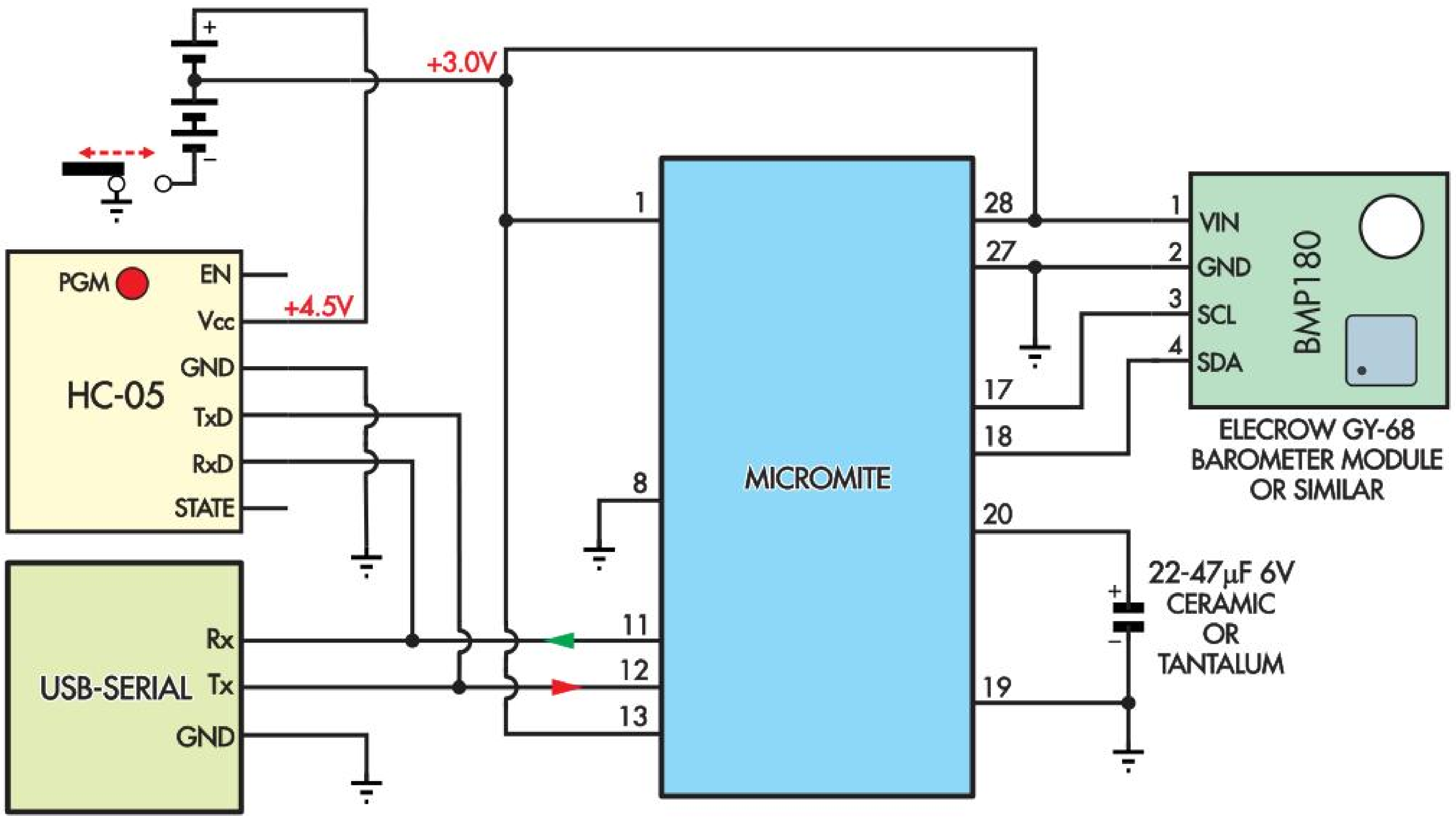
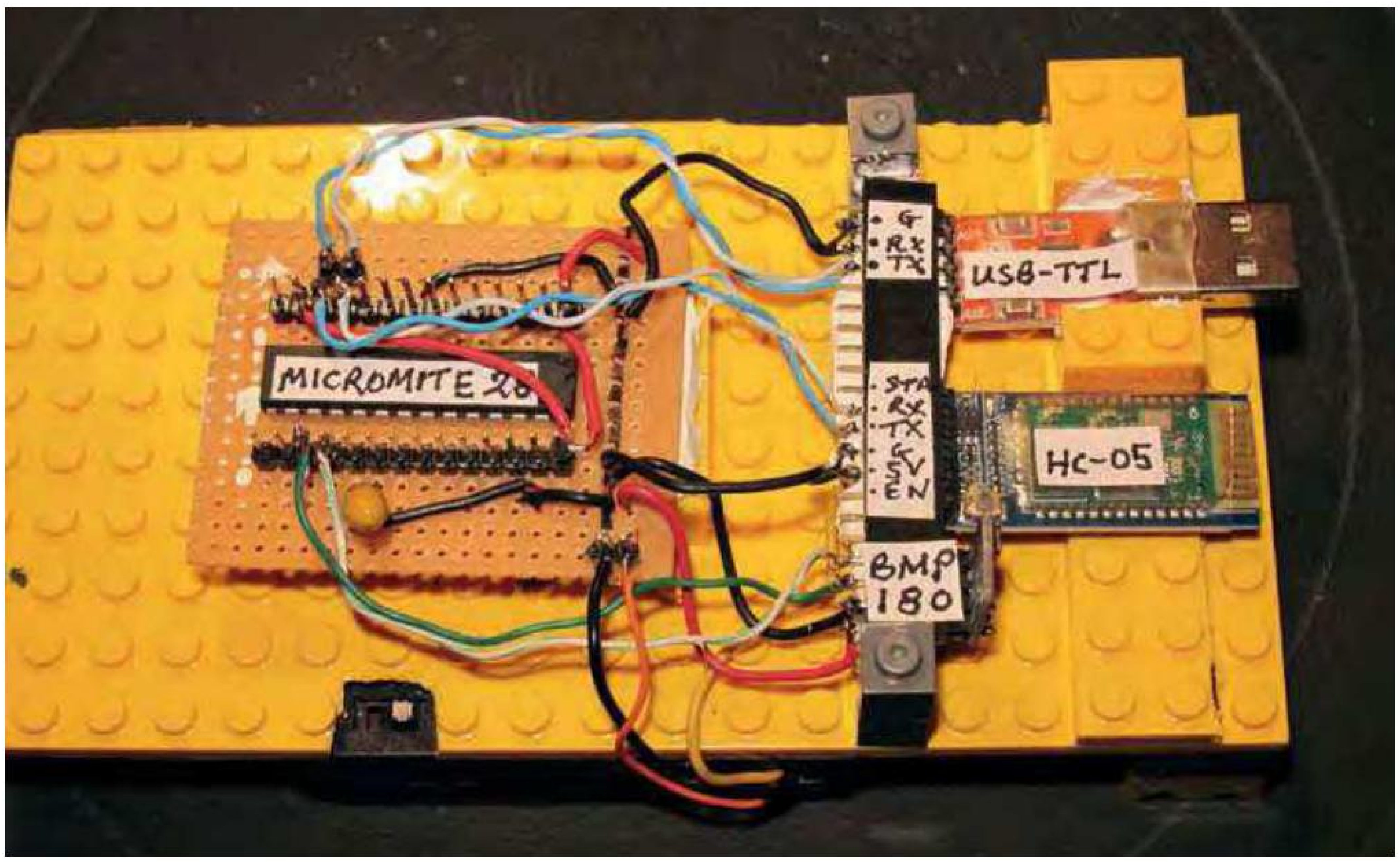

+3.0V
PGM
EN
Vcc
+4.5V
GND
HC-05
TxD
RxD
STATE
1
28
27
17
18
8
MICROMITE
20
1 VIN
2 GND
3 SCL
4 SDA
BMP180
ELECROW GY-68 BAROMETER MODULE OR SIMILAR
22-47µF 6V CERAMIC OR TANTALUM
USB-SERIAL
Rx
Tx
GND
11
12
13
19

Fig.5: by adding a BMP180-based temperature/pressure sensor module as well as the HC-05 Bluetooth module to the Micromite, we can turn it into something useful. It now reports atmospheric data on the smartphone screen via a terminal App.


G
RX
TX
USB-TTL
MICROMITE 28
STA
RX
TX
G
5V
EN
HC-05
BMP 180

The test rig connected to a Micromite Explore-28 which was built on a breadboard. This setup should easily work with a Micromite BackPack and 28-pin Micromite, assuming the requisite pins are free.

While Fig.5 shows both a USB-serial and Bluetooth adaptor, you don't need both; the USB-Serial module is intended mainly for testing and can be left off once you're confident that the HC-05 is working. Also, you don't need to connect the BMP180 module; it's simply there to demonstrate what you can do. Modify the circuit to suit your requirements.

The BMP180 sensor communicates using an I^{2}C serial bus, so it is connected to pins 17 and 18 as shown in Fig.5. It also needs a ground connection and a +3V connection. As before, the 4.5V tap on the battery pack is only required to run the HC-05 module.

Connect the test rig setup to your PC and terminal program via the USB-TTL converter.

We have based the software for this demonstration project (**BMP180 barometer check prog console only.bas**) on the program written by Jim Rowe for the December 2018 article on the GY-68 module with the BMP180 chip (**BMP180 barometer check prog.bas**). Both are available for download from the Spetember 2022 page of the *PE* website, but remeber that here you need to use the '**console only**' version.

The only real change is that all lines which pertain to formatting and/or displaying information on the LCD screen have been removed. Instead, it simply prints the data obtained from the BMP180 chip on the console using PRINT commands.

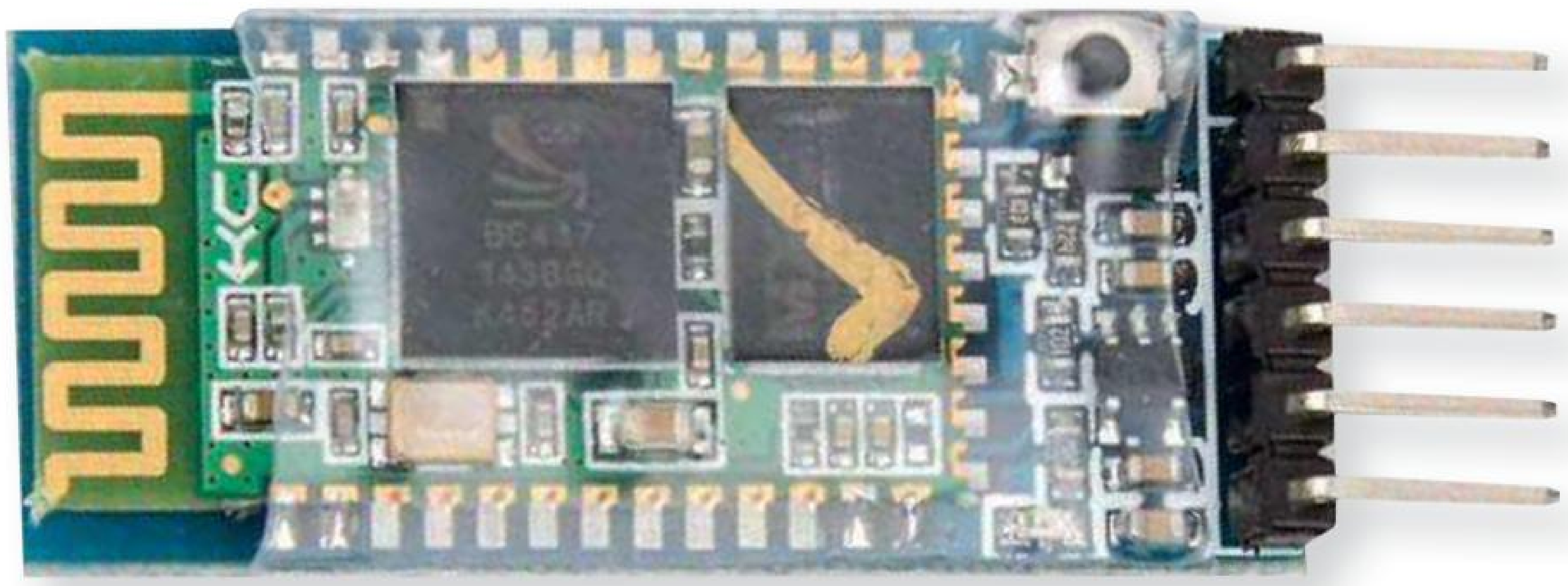
Run the program and confirm that it all performs correctly in the usual PC terminal mode. Then shut down the PC terminal and unplug your test rig from the PC's USB port. Install the Bluetooth Terminal App on your mobile phone (if you haven't already).

Power up the test rig. Notice that the red LED on the HC-05 module is flashing rapidly. Follow the instructions for connecting a Bluetooth device to the Bluetooth Terminal App on your phone.

The steps involve registering the HC-05 in your phone's Bluetooth devices list. It will first show up as an alphanumeric address similar to an IP address but segmented into several pairs of hexadecimal characters. Once you provide the password of 0000 or 1234, your HC-05 should then appear on the list as MMITE01.

Now return to the Bluetooth Serial App on the phone and connect to the MMITE01 adaptor. Successful connection to the HC-05 will be detectable by the flashing LED having slowed down considerably. The App should also display precisely what you have previously seen on your PC's terminal program.

If not, then turn the test rig off and on again. When you turn off the test rig, the Bluetooth Terminal App will report it has lost the connection. Just tap on the connect icon in the App, and it should reconnect without any further need for your inputs or adjustments.



If possible you should try to purchase an HC-05 module which has an 'Enable pushbutton' key, as shown at the upper left of this photo. This specific HC-05 is a HiLetgo branded version.

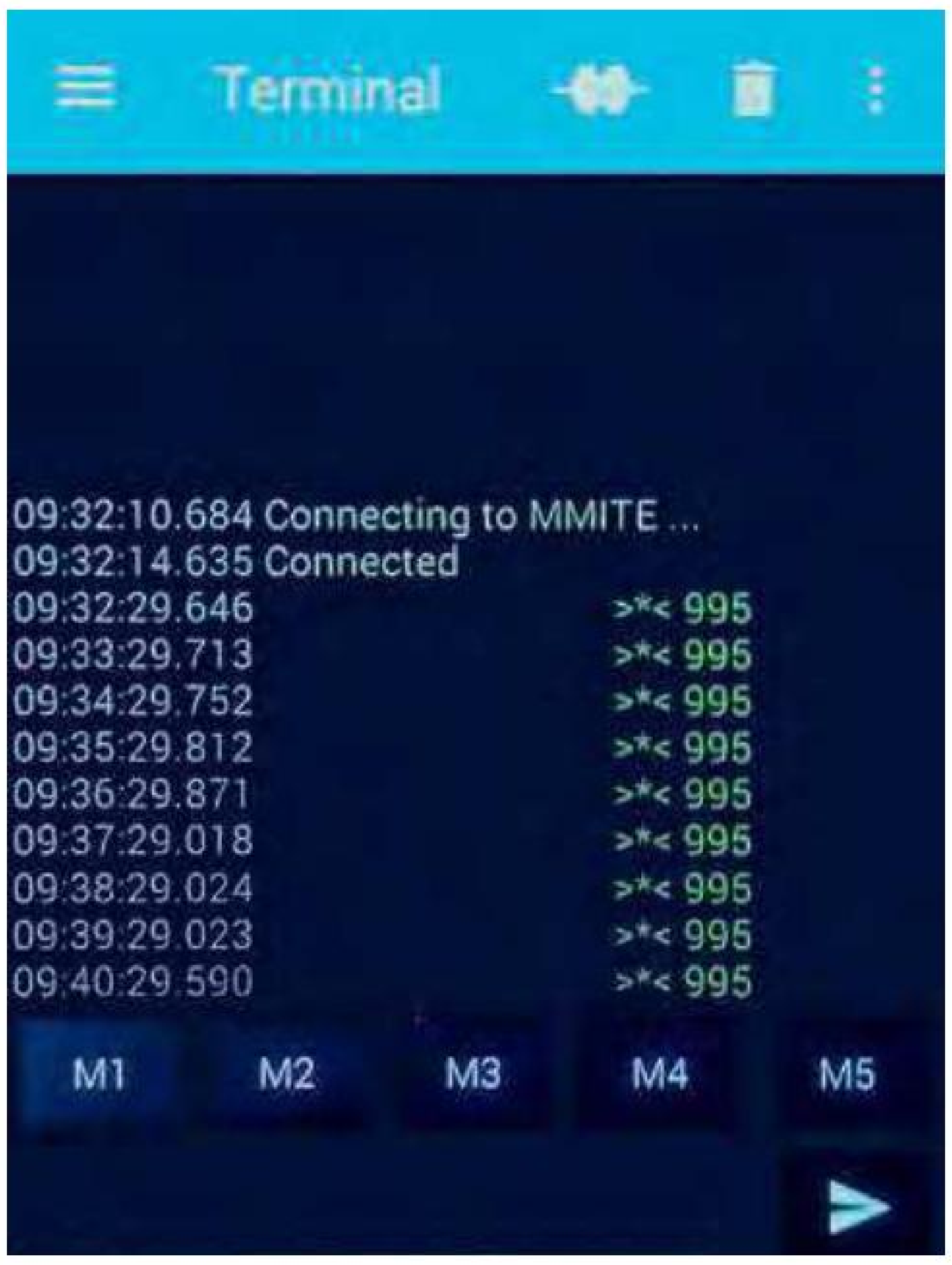
Screen 1 shows a typical display on the mobile phone when connected to the Micromite via Bluetooth. This particular App can log received text, so data coming across from the test rig can be saved.

Another advantage of using this particular Bluetooth Terminal App is that it adds the current date and time to every line of data received, making it unnecessary to build an RTC module into your circuit. In fact, now that the data is in your phone, you can exploit the fact that your phone is, in reality, a very sophisticated computer and display resource.

For example, you can now write your own Android Phone Apps using MIT App Inventor (**ai2.appinventor.mit.edu**) because that tool has a Bluetooth connectivity module as a standard built-in item.

Python programs run well on mobile phones, so that provides another opportunity for enhancing the usefulness of your data collected by the Micromite.

Another possibility is to install a web server on your Android phone, such as KickWeb (**https://bit.ly/pe-sep22-kw**). That way, you can use PHP scripts or continuously looping Python programs to forward sensor derived data to services such as Thingspeak (**www.thingspeak.com**) where your data can be displayed graphically and made available across the whole Internet.



```
09:32:10.684 Connecting to MMITE ...
09:32:14.635 Connected
09:32:29.646                        >*< 995
09:33:29.713                        >*< 995
09:34:29.752                        >*< 995
09:35:29.812                        >*< 995
09:36:29.871                        >*< 995
09:37:29.018                        >*< 995
09:38:29.024                        >*< 995
09:39:29.023                        >*< 995
09:40:29.590                        >*< 995
```

Screen 1: a very basic display of local barometric pressure (in hectopascals [hPa]) in the smartphone terminal app, delivered by the Micromite. By changing the Micromite BASIC code and hardware, you can get it to report just about anything you want!

More Micromite?
Our monthly column *Make it with Micromite* is the best place to learn all about this powerful, easy-to-use single-IC microcontroller.

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HP6644A	PSU 0-60V 3.5A	£400
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HP8341A	Synthesised Sweep Generator 10MHz – 20GHz	£2,000
HP83630A	Synthesised Sweeper 10MHz – 26.5 GHz	POA
HP83624A	Synthesised Sweeper 2 – 20GHz	POA
HP8484A	Power Sensor 0.01-18GHz 3nW-10µW	£75
HP8560E	Spectrum Analyser Synthesised 30Hz – 2.9GHz	£1,750
HP8563A	Spectrum Analyser Synthesised 9kHz – 22GHz	£2,250
HP8566B	Spectrum Analsyer 100Hz – 22GHz	£1,200
HP8662A	RF Generator 10kHz – 1280MHz	£750
Marconi 2022E	Synthesised AM/FM Signal Generator 10kHz – 1.01GHz	£325
Marconi 2024	Synthesised Signal Generator 9kHz – 2.4GHz	£800
Marconi 2030	Synthesised Signal Generator 10kHz – 1.35GHz	£750
Marconi 2023A	Signal Generator 9kHz – 1.2GHz	£700

Marconi 2305	Modulation Meter	£250
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Marconi 2945/A/B	Communications Test Set Various Options	POA
Marconi 2955	Radio Communications Test Set	£595
Marconi 2955A	Radio Communications Test Set	£725
Marconi 2955B	Radio Communications Test Set	£800
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Marconi 6960B	Power Meter with 6910 sensor	£295
Tektronix TDS3052B	Oscilloscope 500MHz 2.5GS/s	£1,250
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Tektronix 2430A	Oscilloscope Dual Trace 150MHz 100MS/s	£350
Tektronix 2465B	Oscilloscope 4 Channel 400MHz	£600
Farnell AP60/50	PSU 0-60V 0-50A 1kW Switch Mode	£300
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Farnell AP100-90	Power Supply 100V 90A	£900
Farnell LF1	Sine/Sq Oscillator 10Hz – 1MHz	£45
Racal 1991	Counter/Timer 160MHz 9 Digit	£150
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Racal 9300	True RMS Millivoltmeter 5Hz – 20MHz etc	£45
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Solartron 7150/PLUS	6½ Digit DMM True RMS IEEE	£65/£75
Solatron 1253	Gain Phase Analyser 1mHz – 20kHz	£600
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Tasakago TM035-2	PSU 0-35V 0-2A 2 Meters	£30
Thurlby PL320QMD	PSU 0-30V 0-2A Twice	£160 – £200
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HP/Agilent HP 34401A Digital
Multimeter 6½ Digit £325 – £375

HP33120A	Function Generator 100 microHz – 15MHz	£350
HP53131A	Universal Counter 3GHz Boxed unused	£600
HP53131A	Universal Counter 225MHz	£350
Audio Precision	SYS2712 Audio Analyser – in original box	POA
Datron 4708	Autocal Multifunction Standard	POA
Druck DPI 515	Pressure Calibrator/Controller	£400
Datron 1081	Autocal Standards Multimeter	POA
ENI 325LA	RF Power Amplifier 250kHz – 150MHz 25W 50dB	POA
Keithley 228	Voltage/Current Source	POA
Time 9818	DC Current & Voltage Calibrator	POA

Marconi 2955B Radio
Communications Test Set – £800

BY JOHN CLARKE

20A DC Motor Speed Controller

Feedback
IC2
Soft Start
CON2
TP1
Hysteresis
TP4
TP5
TP6
Under-
Max
Q1
Q3
0V
V+
Motor Supply
Supply
M+
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This small but powerful speed controller has a 20A rating and is packed with features. It suits a wide range of applications and is simple to build and use. Features include low-battery protection, soft starting and adjustable pulse frequency. It can handle DC motors that run from near 0V up to 30V.

There are a great many applications for DC motors where speed control is wanted or necessary. Since DC motors can be run directly from batteries, they are used in golf carts, electric scooters, bikes, skateboards, remote-controlled cars and boats – the list goes on.

In most of those applications, you need a way to control the speed of the motor. Going flat out all the time isn't always a good idea!

A speed controller like this one is the ideal solution. It can handle DC motors with a rated voltage of up to 24V (30V maximum) and continuous currents up to 20A.

This controller is presented as a bare electronic module built on a PCB that can be installed in a standard UB3 plastic case. It includes heavy-duty terminals for the power supply and motor connections, plus additional terminals for the speed control potentiometer that mounts off the PCB.

The motor-driving components are mounted on substantial heatsinks for cooling. The adjustable features like soft-start rate and feedback gain are set using onboard multi-turn trimpots with voltage test points. An onboard LED indicates the speed setting, as well as faults like low battery or motor disconnection.

Speed controller design

While we have published many DC motor speed controllers in the past, this version has more features and better performance. The motor speed is controlled using pulse-width modulation (PWM). That means that the motor is driven by a series of on and off voltage pulses rather than a variable DC supply, making it more efficient.

Speed control of the motor is done by varying the pulse width. The ratio of the pulse width to the interval between pulses is the duty cycle. A low duty cycle will only provide a voltage to the motor for a small portion of the time, and the motor runs slowly.

As the on-pulse duration increases, this greater duty cycle makes the motor run faster until it reaches 100% duty cycle and is driven continuously.

Oscilloscope traces Scope 1 and Scope 2 show how this PWM scheme works. In Scope 1, the top (yellow) trace is the gate drive signal for MOSFETs Q1 and Q2. When it is high, the motor is powered. In this case, the duty cycle is very low at about 9.5%, so the motor runs slowly. The lower cyan trace is related to the motor current. This is used to maintain motor speed with a variable load.

Scope 2 has the same two traces, but this time the duty cycle is much higher, and the motor runs faster. The motor is loaded less than in Scope 1, so the current reading is lower despite the higher duty cycle.

What's new

One of the problems with controlling DC motors using PWM is that the motor can make extra noise due to the motor windings and other mechanical parts vibrating at the PWM frequency. This can be alleviated to some extent by adjusting the PWM frequency to produce minimal noise.

That noise tends to be reduced as the PWM frequency is increased, and is mostly eliminated at PWM frequencies above 20kHz (around the upper limit of human hearing).

But increasing the frequency can cause problems too. It becomes harder to maintain the motor speed against a varying motor load using the traditional back-EMF feedback system. Very high PWM frequencies can also cause a loss of motor torque.

These problems and solutions are described in more detail in the section titled, *PWM motor driving pitfalls at higher frequencies*.

This controller gives you the ability to adjust the PWM frequency beyond audibility while addressing the problems of limited low-speed motor torque and control at elevated frequencies.

Other features that are incorporated include soft starting, low-voltage cutout, LED status indication and optional motor disconnected detection. These features are easy to set up and adjust via trimpot adjustments.

Soft starting

This is where the motor is slowly increased in speed, up to the setting of the speed pot. Soft starting reduces the surge of current and a rapid build-up of motor torque compared to applying power suddenly. The PWM duty cycle is ramped up over a longer period, so that the motor starts more smoothly.

The maximum soft-start period is two seconds for the full range from 0% duty to 100%. This period can be adjusted from zero to two seconds in 255 steps.

Soft starting can be initiated in several ways. It applies when the controller is initially powered up, or when the speed control is started from the fully off position, and finally, after returning to regular operation from low-voltage shutdown.

Features

- DC motor PWM drive
- Can drive motors rated up to 24V and 20A DC
- Motor and controller supply voltage can be separate
- 16 PWM frequency choices
- Motor load feedback control and gain adjustment
- Adjustable soft-start rate
- Motor speed curve adjustment
- Under-voltage cut-out with LED indication and adjustable hysteresis
- Duty cycle LED indicator
- Optional motor disconnect detection

Specifications

- Speed adjustment range: 0% to 100% duty cycle
- Motor supply: from near-zero to 30V maximum
- Controller supply: 10.5V to 30V maximum (5.5-26V with ZD1 linked out)
- Speed indication: LED1 brightness varies with PWM duty cycle
- PWM frequency: 16 steps from 30.6Hz to 32.4kHz (see Table 1)
- Soft-start rate: 0-2 seconds in 255 steps for 0% to 100% duty cycle
- Speed curve adjustment: minimum speed can be set to 0-33% duty cycle
- Under-voltage (UV) threshold: 0-30V in 29.6mV steps
- UV hysteresis: 0-5V in 29.6mV steps
- UV indication: LED1 flashes on for 65ms at 1Hz
- Motor disconnection detection: motor is shut down if monitored current drops to zero while driving motor; indicated with 2Hz/50% duty cycle LED flashing
- Speed pot disconnection detection: indicated with a dimly illuminated LED

Low-voltage detection

The low-voltage detection feature is included to prevent over-discharging a battery supplying power to the motor. Most batteries, including lead-acid and lithium chemistry types, will be damaged if discharged beyond a certain voltage.

This feature switches off the motor drive at a pre-set threshold voltage. This is indicated with a 65ms flash of the indicator LED at 1Hz.

The voltage must be below the threshold for more than ten seconds before the drive to the motor is switched off. This prevents any nuisance low-voltage trips that would otherwise switch off the controller due to a short-term voltage drop when the motor starts up.

Once shut down, the voltage needs to rise above the low-voltage detection threshold by a certain amount before it will start up again. This hysteresis prevents constant switching on and off as the battery voltage recovers with the motor load removed, only to switch off again once the motor restarts.

Motor disconnection

The optional motor disconnect detection prevents the motor from starting up if it is disconnected and then reconnected while the speed setting is above zero. When the motor is detected as disconnected, the speed potentiometer needs to be wound fully anticlockwise and the motor reconnected before it can run again. The disconnected state is indicated with the indication LED blinking at 2Hz.

Separate supply voltage

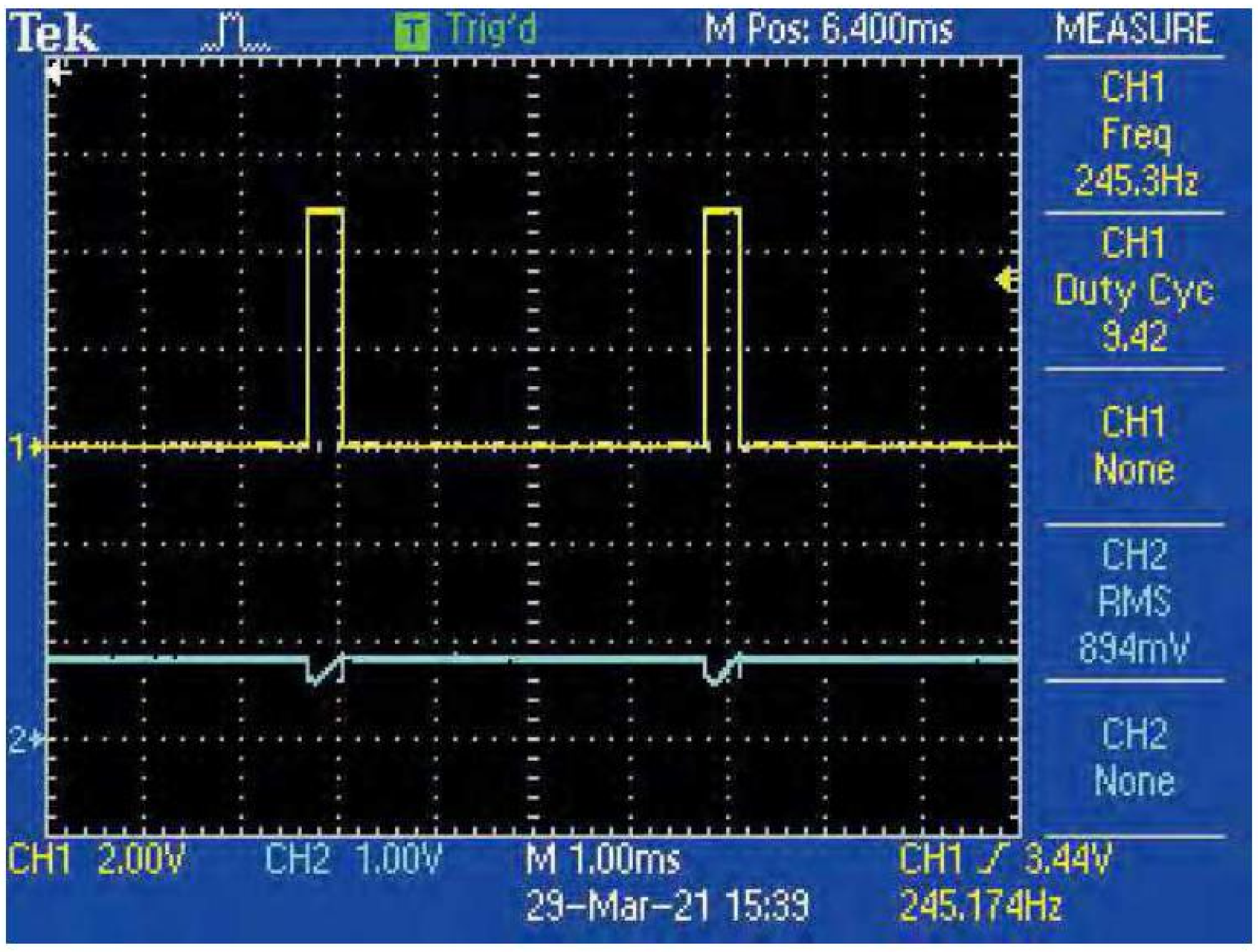
Another feature is the ability to separate the controller's supply voltage from the supply to the motor. This means that the motor can be run from a much lower supply voltage than that required to operate the *DC Motor Speed Controller*.

So while the *DC Motor Speed Controller* requires a supply of at least 10.5V to operate (up to 30V), the motor can be run using a separate supply from near 0V up to 30V. The 30V limit is sufficient to allow for just about any 24V battery; eg, a fully charged 12-cell lead-acid battery is around 29V.

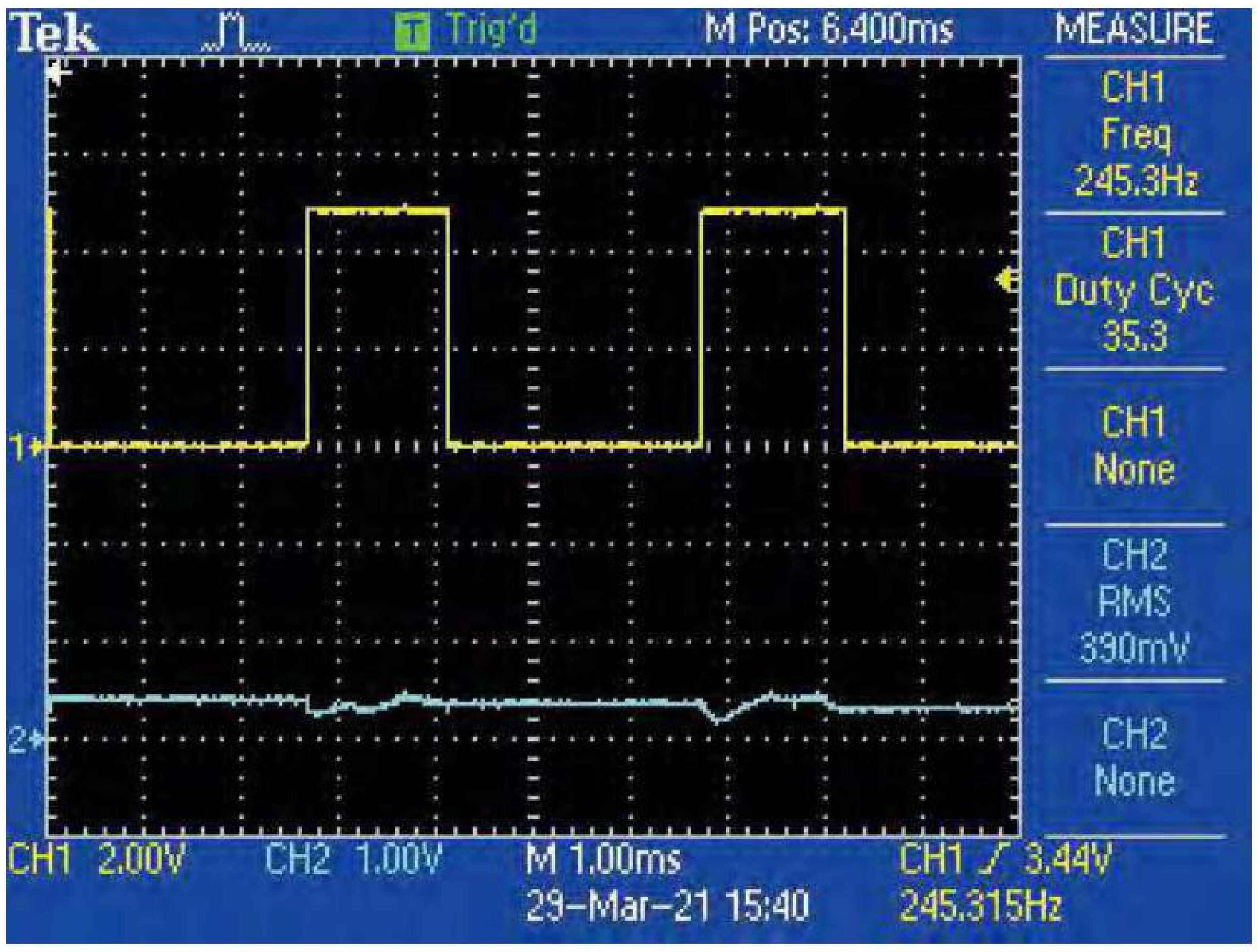
You can use the same supply for both the controller and the motor, provided the voltage is in the 10.5-30V range, and that voltage is suitable for the motor.

Circuit details

The *DC Motor Speed Controller* circuit is shown in Fig.1. It is based around an 8-bit PIC16F1459 microcontroller (IC1) which provides the PWM drive signal and monitors the battery voltage, motor current and the voltage from several trimpots and the speed potentiometer. IC1 also monitors rotary switch S1, which selects the PWM frequency.



Scope 1: a pulse-width modulated (PWM) drive signal at a low duty cycle, about 9.5%. Current has little time to build during each pulse, so the motor runs slowly.



Scope 2: another PWM drive signal, here with a duty cycle of 35.5%, which is roughly the same as driving the motor at 1/3 supply voltage, so it will run faster but not nearly at full speed.

IC1 has two PWM outputs, and we use both. One is at pin 5 (PWM1) and the other at pin 8 (PWM2). These PWM outputs have different functions, but provide the same PWM frequency and duty cycle most of the time while the motor is being driven.

The PWM1 output is used to drive MOSFETs Q1 and Q2 via gate driver IC3. IC3 is an MCP1416, designed to provide a high-current drive with fast rise and fall times to the MOSFET gates. This ensures that they switch on and off quickly. Each MOSFET gate is isolated from the other using a 10Ω resistor. The resistors also prevent MOSFET switching oscillations at the gate threshold.

These MOSFETs are logic-level types that fully conduct with a gate voltage of 5V. Non-logic-level MOSFETs typically require at least 10V for full conduction. The two MOSFETs are connected in parallel, sharing the load (motor) current.

Low-value resistors are placed between the source of each MOSFET and ground, with Q1's source resistor being used to monitor the current. The source resistor on Q2, while not used for load current measurement, is still necessary. That's so that the total on-resistance of MOSFET Q2 and its source resistor matches Q1 and its source resistor.

Since the MOSFET on-resistance is typically 0.014Ω, the 0.01Ω source resistor for Q2 helps maintain even sharing of the load current between the two MOSFETs. Without it, Q2 would carry about 2/3 of the load current and Q1 only 1/3.

Diode D1 is included between the positive supply and the MOSFET drains to clamp the induced voltage spike when the motor's drive is switched off. This diode is effectively connected across the motor terminals. It is a dual 10A schottky diode that can conduct 20A continuously when the diodes are connected in parallel. Paralleling the diodes ensures nearly equal current sharing. That is possible because the two diodes are on the same silicon die, and therefore have the same characteristics and operating temperature.

The motor supply is connected to the GND and motor supply + terminals on screw connector CON1. This positive supply is fed to the motor via fuse F1, an automotive blade-type fuse with a rating selected to suit the motor. Three 470µF 35V low-ESR electrolytic capacitors bypass the motor supply after the fuse. These are to provide a high short-term peak current supply.

Feedback control

Many DC motor speed controllers monitor motor back-EMF (electromotive force) to determine when variations in the load might reduce the speed of the motor. This back-EMF is the voltage generated by the motor when the supply to it is switched off and the motor is still turning. The induced voltage reduces when the motor slows under load.

Speed control is maintained by increasing the PWM duty cycle to increase motor torque and speed when its speed drops. But we don't use the back-EMF sensing method for reasons described under the section *PWM motor driving pitfalls at higher frequencies*. Instead, we monitor its current draw.

When MOSFETs Q1 and Q2 are conducting, the voltage across Q1's 0.01Ω source resistor is proportional to the current being drawn by the motor. When the MOSFET is off, there is no voltage across this resistor. So we use a sample-and-hold circuit to capture the voltage while Q1 is conducting.

MOSFET Q3 and the 100µF capacitor form the sample-and-hold buffer. The gate of Q3 is driven by the PWM2 output of IC1, which follows the PWM1 output. So when Q1 and Q2 are on, so is Q3, and the 100µF capacitor charges or discharges so that its voltage approaches that across the 0.01Ω current-sense resistor.

When MOSFETs Q1 and Q2 switch off, so does Q3, isolating the 100µF capacitor from the 0.01Ω resistor to prevent it discharging during the off-time.

The reason we use the separate PWM2 output to drive Q3 has to do with the case when the motor is off. In this case, the PWM1 output has a duty cycle of 0% (ie, it's held low), but PWM2 is programmed to produce a 60µs pulse every 13.4s. This switches Q3 on momentarily, discharging the 100µF capacitor via the 0.01Ω resistor.

This on-duration is extended if the capacitor needs to be discharged from a higher voltage, especially when the motor is turned off by reducing the speed control. Without this, the 100µF capacitor slowly charges via leakage current from amplifier IC2, causing the motor to start rather abruptly.

IC2 is an instrumentation amplifier and provides amplification of the small voltage across the shunt for current measurement. Its gain can be adjusted from between 611, when trimpot VR6 is at minimum resistance, and about nine when the trimpot is at its maximum of 50kΩ. This caters for the wide range of motors that could be used, ranging from those drawing less than 1A up to 20A.

The output from IC2 is monitored by the AN9 analog input (pin 9) of microcontroller IC1, which uses its internal analogue-to-digital converter (ADC) to convert the voltage from IC2 into a 10-bit digital value (0 to 1023).

Speed control

Potentiometer VR1 is the main speed control. The voltage at its wiper varies with its rotation, and is fed to analog input AN5 (pin 15) of IC1. This is converted to a 10-bit digital value, indirectly controlling the PWM duty cycle applied to the MOSFETs.

Motor load compensation is performed by increasing the duty cycle of the PWM signal depending on the motor load, based on the motor current. The amount of feedback applied is adjusted by setting the gain for IC2, as described above.

Supply voltage monitoring

The motor supply voltage is monitored at analog input AN10 (pin 13) of IC1. The supply voltage is reduced to one-sixth (1/6) of its full value by a 10kΩ/2kΩ voltage divider. So for a 0-30V motor supply, the voltage at AN10 is in the range 0-5V.

This voltage is filtered using a 100nF capacitor to prevent noise from altering the result of the ADC conversion.

Setting adjustments

This voltage is compared with the under-voltage threshold setting voltage at the AN7 input (pin 7) set by trimpot VR4. This trimpot is connected across the 5V supply, allowing a voltage range adjustment from 0-5V. Test point TP4 is included so the set threshold can be measured.

To make setting up easier, the voltage at TP4 is one-tenth the under-voltage threshold. So if you want the under-voltage threshold to be 11.5V, set the voltage at TP4 to 1.15V.

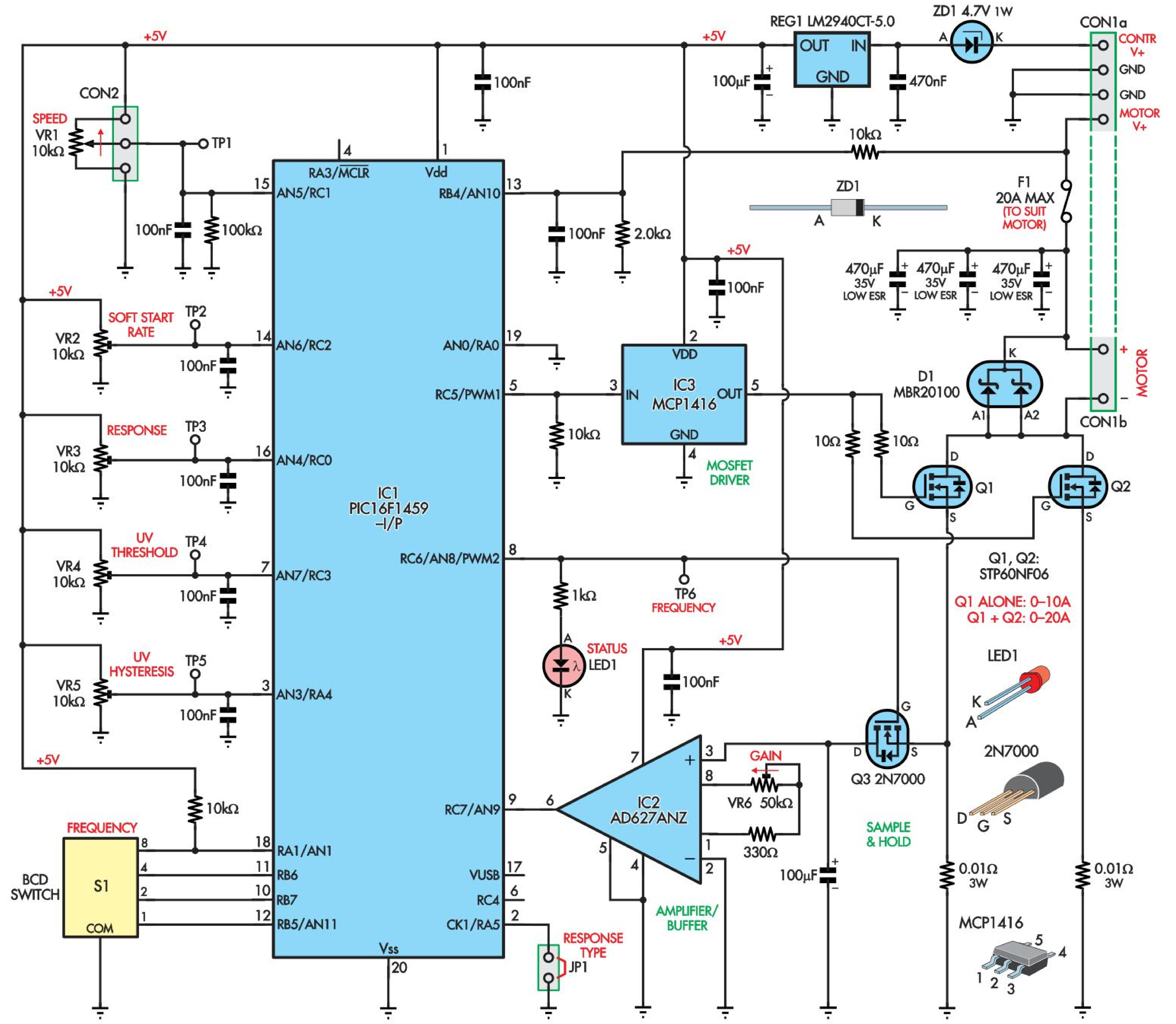
The voltage at the AN7 input is converted to a digital value and multiplied by 1.6666, so the scale matches the divided-by-six motor voltage.

The motor supply has to drop below this threshold for 10 seconds before the drive to the motor is switched off. When this happens, LED1 flashes momentarily each second.

Typically, a battery will recover a little when the motor drive is switched off; the battery voltage will rise once there is no load. To prevent the motor from switching on again due to this effect, we add hysteresis.

The motor supply will need to go above the low voltage threshold plus the hysteresis voltage before the motor drive will be re-enabled. In practice, the battery needs to be charged before the motor can run again.

This hysteresis is set using trimpot VR5, monitored at TP5. The TP5 reading is the full hysteresis voltage (not 1/10th as for the threshold measurement at TP4). So if you want a 1V hysteresis, adjust VR5 until TP5 reads 1V.


+5V
CON2
SPEED
VR1
10kΩ
TP1
100nF
100kΩ
100nF
REG1 LM2940CT-5.0
OUT
IN
GND
100µF
470nF
ZD1 4.7V 1W
A
K
CON1a
CONTR V+
GND
GND
MOTOR V+
10kΩ
ZD1
F1
20A MAX
(TO SUIT MOTOR)
2.0kΩ
470µF 35V LOW ESR
470µF 35V LOW ESR
470µF 35V LOW ESR
RA3/MCLR
Vdd
AN5/RC1
RB4/AN10
AN6/RC2
AN0/RA0
RC5/PWM1
AN4/RC0
IC1
PIC16F1459
–I/P
AN7/RC3
RC6/AN8/PWM2
AN3/RA4
RC7/AN9
RA1/AN1
RB6
RB7
RB5/AN11
VUSB
RC4
CK1/RA5
Vss
SOFT START RATE
TP2
VR2
10kΩ
RESPONSE
TP3
VR3
10kΩ
UV THRESHOLD
TP4
VR4
10kΩ
UV HYSTERESIS
TP5
VR5
10kΩ
FREQUENCY
BCD SWITCH
S1
COM
VDD
IN
IC3
MCP1416
OUT
GND
MOSFET DRIVER
10kΩ
1kΩ
TP6
FREQUENCY
STATUS LED1
D1
MBR20100
A1
A2
10Ω
10Ω
Q1
Q2
Q1, Q2: STP60NF06
Q1 ALONE: 0–10A
Q1 + Q2: 0–20A
+
–
MOTOR
CON1b
LED1
2N7000
D G S
Q3 2N7000
GAIN
VR6 50kΩ
330Ω
IC2
AD627ANZ
AMPLIFIER/ BUFFER
100µF
SAMPLE & HOLD
0.01Ω 3W
0.01Ω 3W
MCP1416
RESPONSE TYPE
JP1

20A DC Motor Speed Controller

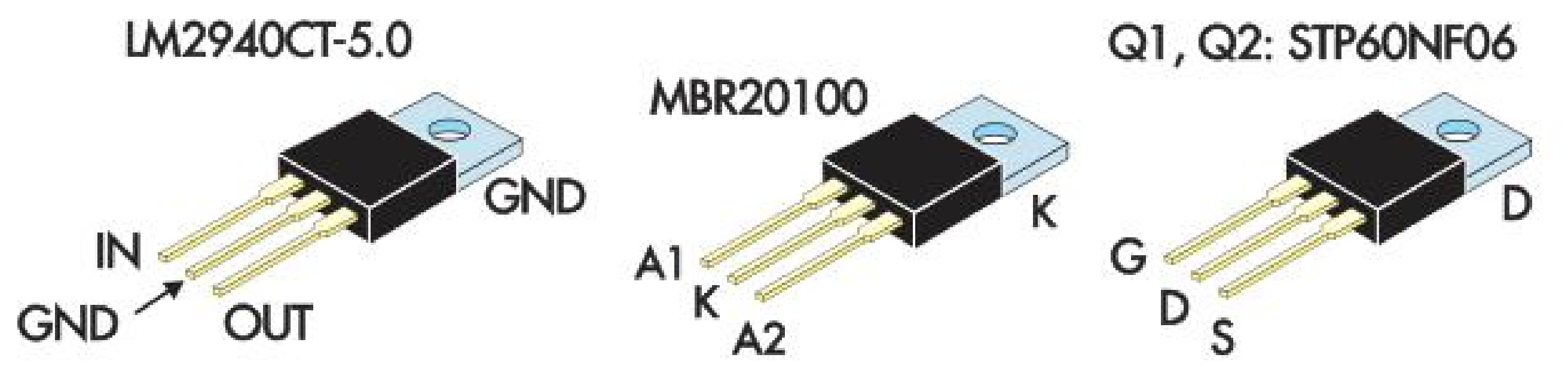

LM2940CT-5.0
IN
GND
OUT
GND
MBR20100
A1
K
A2
K
Q1, Q2: STP60NF06
G
D
S
D

Fig.1: microcontroller IC1 monitors the positions of speed pot VR1 and trimpots VR2-VR5 via five analogue input pins. It also reads the position of BCD switch S1 (used to set the PWM frequency) using four digital inputs. A PWM waveform is produced at pin 5, which drives MOSFETs Q1 and Q2 via driver IC3; those MOSFETs switch current through the motor. The motor current is converted to a voltage using a 10mΩ shunt; this voltage is amplified by IC2 and measured at pin 9 of IC1.

The soft-start period adjustment is with VR2, measured at TP2. This voltage is monitored at the AN6 input, and sets the maximum rate at which the motor speed increases.

The maximum time to reach 100% duty cycle from zero is two seconds, with 5V at TP2. A 2.5V setting will give a one-second soft-start period, and so on.

VR3 is the speed curve adjustment trimpot, with corresponding test point TP3. This is monitored at the AN4 input of IC1, pin 16. This allows the speed pot to be used over its entire range when the PWM frequency is set relatively high, and can also compensate for the fact that motors can require a duty cycle well above 0% before they start spinning.

As described in the separate panel labelled, *PWM motor driving pitfalls at higher frequencies*, in some cases, driving a motor with a high PWM frequency can mean that the motor will not start until the duty cycle is at 20%, or even higher.

The curve adjustment sets the initial duty cycle when the speed potentiometer is rotated just clockwise from fully-anticlockwise. This adjustment removes the dead zone from the speed pot. The curve adjustment range is from almost zero to a 33% initial duty cycle.

Whenever the curve setting is non-zero, the software within IC1 expands the speed control range so that the maximum duty cycle is still achieved when VR1 is fully clockwise.

Operation at low frequencies can also be optimised using the curve adjustment, with jumper JP1 inserted to pull the normally high RA5 digital input low (pin 2). Without the jumper inserted, the RA5 input is pulled high via an internal pull-up current.

The curve adjustment when JP1 is inserted allows for better feedback control at very low duty cycles. The adjustment reduces the motor snap-on effect, where the feedback voltage suddenly rises with an increase of the PWM duty just off from zero. This adjustment sets a feedback offset value

PWM motor driving pitfalls at higher frequencies

When using PWM to drive a DC motor, the average motor winding current varies depending upon the duty cycle. Since torque is proportional to the winding current, the motor speed can be easily controlled.

In theory, the motor speed is not affected by the frequency; it is only the duty cycle that matters because that sets the average current through the motor windings. Higher PWM frequencies will result in less ripple in the motor current, but will not affect the average significantly.

But there are cases where higher frequencies can affect the current at lower duty cycles, to the point that the motor will refuse to turn at all with lower duty cycles. There is much confusion over the reasons for this and what to do about it.

We trawled the internet trying to find a good explanation of this phenomenon, and most of the information we came up with was misleading or incorrect. So we performed several experiments to find out for ourselves.

We found: if you are using a half-bridge or full-bridge to drive a DC motor, it will behave pretty much as theory predicts. The motor current varies almost exactly linearly with the PWM duty cycle, regardless of frequency.

That is what you would expect if you model the motor as an inductance in series with a resistance. If the inductance is L and the series resistance is R, the motor winding impedance is then $R + j2\pi fL$. The magnitude of the current for a sinewave at any given frequency f is then: $V \div \sqrt{(R^2 + (2\pi fL)^2)}$.

A PWM signal comprises a DC component (the average level, V × duty cycle) plus AC components at the switching frequency f, and its squarewave harmonics at 3f, 5f, 7f... The exact mix of harmonics varies with the duty cycle.

As the current decreases with increasing frequency, the winding inductance attenuates the AC components of the PWM signal. The motor windings act to smooth out these ripples, but the inductance has no effect on the direct current level; it is solely determined by the supply voltage, duty cycle and motor winding resistance.

Our tests bear this out. But like many simpler designs, our motor speed controller does not use a half-bridge or full-bridge design and therefore does not produce a square wave across the motor windings.

The motor's positive terminal is connected to V+, and the negative end is periodically pulled down to 0V when MOSFETs Q1 and Q2 switch on.

Some of the time, we have V+ across the motor. But the rest of the time, when MOSFETs Q1 and Q2 are off, the winding inductance and back-EMF pull the motor's negative terminal above the positive terminal. The voltage is clamped by diode D1 to around 0.5V above the positive voltage.

So there is a negative voltage across the motor when the MOSFETs are off, rather than 0V, and a significant recirculating current flows through diode D1. This causes the motor winding current to decay significantly faster than in the half-bridge or full-bridge case described above.

You can see this if you compare Scopes 3 and 4. These show the same unloaded DC motor being driven at the same PWM frequency (3.92kHz) and same duty cycle (10%) but with half-bridge drive in Scope 3 and single-ended drive in Scope 4. The yellow trace shows the applied voltage, while the green trace shows the current through the motor windings.

The rate of current rise and peak current are similar between the two. But when the high-side MOSFETs switch off and the low-side MOSFETs switch on in Scope 3, you can see an exponential decay in the motor winding current. The current flows throughout the whole cycle until it starts rising again on the next cycle.

In Scope 4, with the current recirculating through the diode during the off-time, it decays exponentially (but faster), then linearly, reaching zero before the next cycle. Therefore, the average current is much lower, around half (a reading of 400mV vs 800mV), despite the duty cycle being the same.

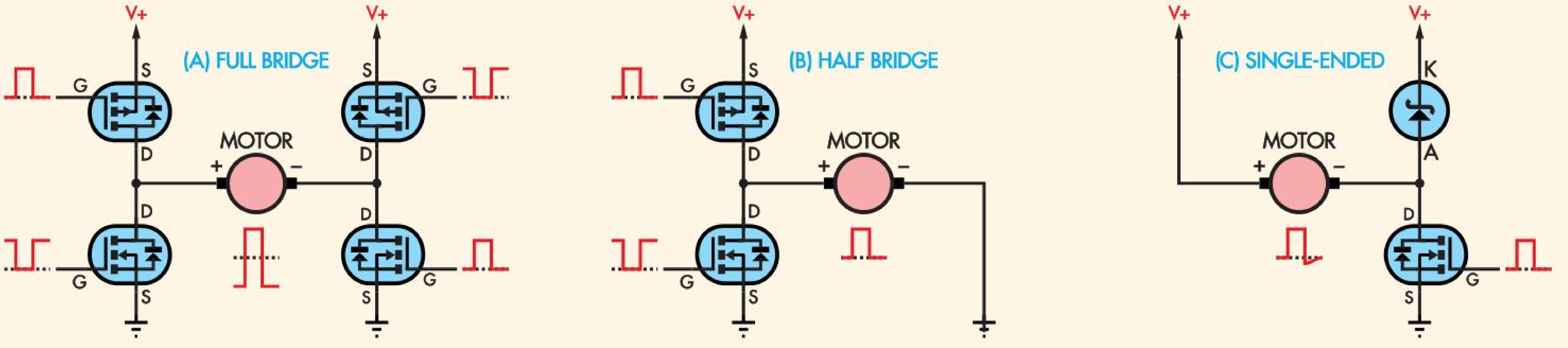
Scope 5 shows the same half-bridge drive scheme used in Scope 3, again with a 10% duty cycle, but at a much higher PWM frequency of 31.4kHz. The average current is only a little bit lower, reading about 750mV compared to around 800mV, due to the MOSFET 'dead time' being more significant at this higher switching frequency.

Scope 6 shows the same single-ended drive scheme as in Scope 4, but this time at 31.4kHz. The current disparity has increased further – the average winding current is now only 286mV. So the effect of the single-ended drive scheme on motor current is worse at higher frequencies.

With the single-ended drive scheme, the average motor current for low duty cycles is less than expected, and this effect increases at higher frequencies. So it is a good idea to increase the minimum duty cycle at higher PWM frequencies to compensate, which is the reason for trimpot VR3 in this design.

The magnitude of this effect can vary with the motor, too. Larger motors with a higher inductance will tend to suffer more from reduced current (and torque) at low duty cycles with higher PWM frequencies.

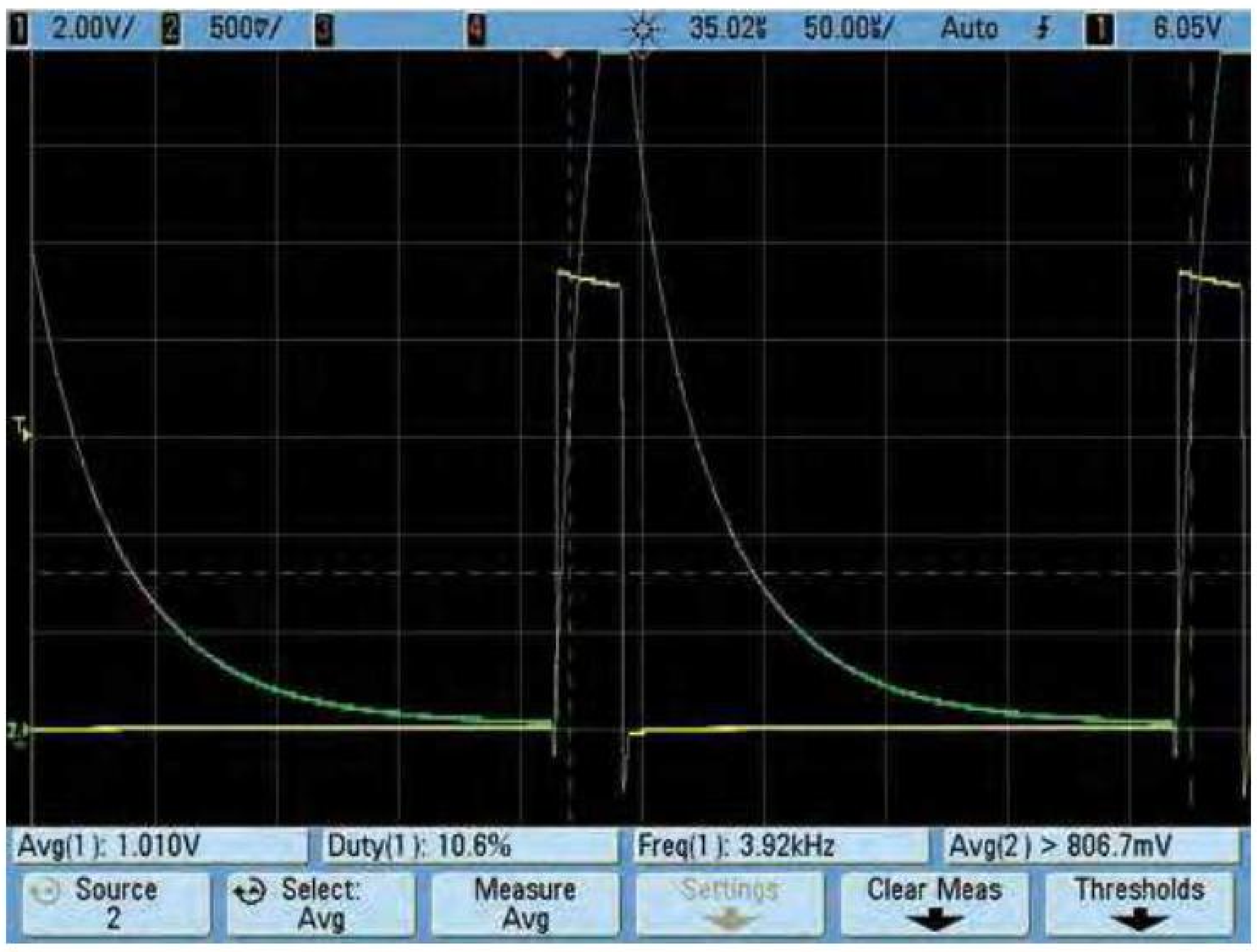
In practice, the easiest way to compensate for this effect is to tune the minimum duty cycle setting (by adjusting VR3) until you get satisfactory speed control at the lower end of speed pot VR1's range. If this cannot be achieved for a given motor, try a lower PWM frequency.



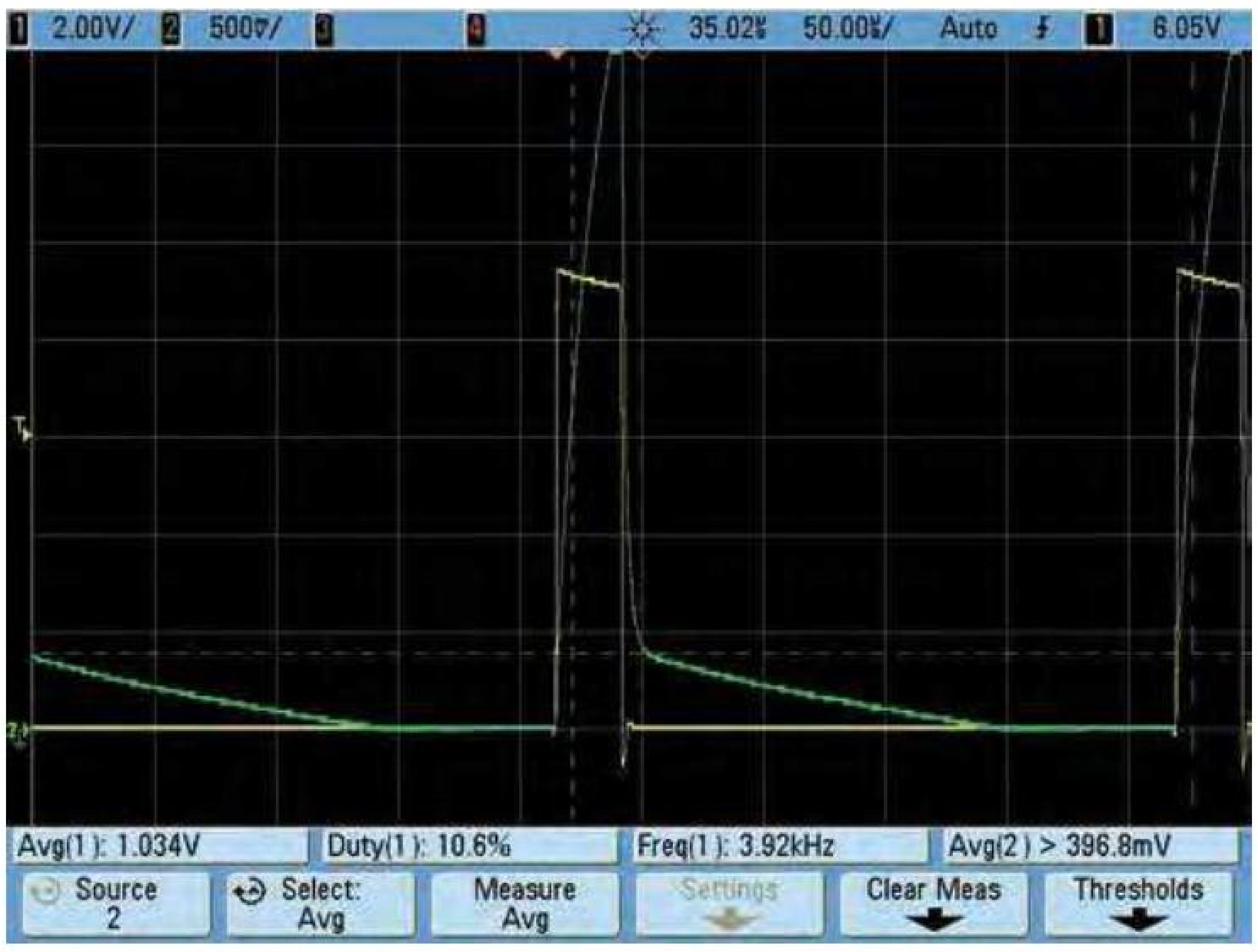
so that feedback is ignored below the specified speed setting.

Trimpot VR3 is also used to enable or disable motor disconnection detection. This is done by splitting VR3's range into two halves, 0-2.5V and 2.5-5V. From 0V to 2.5V, motor disconnection checking is disabled. Above 2.5V, motor disconnect detection is enabled and the curve adjustment is reversed, with fully clockwise giving the same effect as fully anti-clockwise.

When the motor current feedback is below a set value for more than about 200ms, the motor is determined as being



Scope 3: the voltage across the motor (yellow) and current (green) with a half-bridge at 10% duty cycle. The motor inductance limits the current rise and fall times. The current does not fall back to zero before the next pulse, despite the relatively low duty cycle; the winding inductance sustains it.



Scope 4: like Scope 3, but we have switched from a half-bridge driver to a single MOSFET with a recirculating diode, as used in this (and many other) speed controllers. This dramatically affects how the current tapers off at the end of each pulse, so the motor current is much lower with low duty cycles.

disconnected. In this case, the PWM duty cycle is set to zero and the LED flashes at 2Hz.

The motor will only start again after it is reconnected, and the speed pot is wound fully anti-clockwise. This prevents erratic operation due to, for example, loose wires.

Motor disconnect detection is optional, because unless the motor is set up correctly when used at high frequencies, false disconnection events can cause nuisance shutdowns. This can occur if the curve is not adjusted correctly, with a sufficiently high duty cycle at the start of the speed pot rotation.

PWM frequency options

Switch S1 is used to select the frequency of the PWM drive for the motor. This is a 16-position rotary BCD (binary-coded decimal) switch. There are four switch terminals labelled 8, 4, 2 and 1, plus a common connection, which we have connected to ground.

The other switch terminals connect to the RA1, RB6, RB7 and RB5 digital inputs of IC1, respectively. All of these pins except for RA1 are configured in IC1 to provide a pull-up current. The RA1 input does not have such an option, so an external 10kΩ pull-up resistor connects to 5V.

These pull-ups hold the inputs high (at 5V) whenever the switch does not connect that terminal to ground. The 16 possible combinations are decoded in IC1, and the required PWM frequency is selected (see Table 1).

Power supply

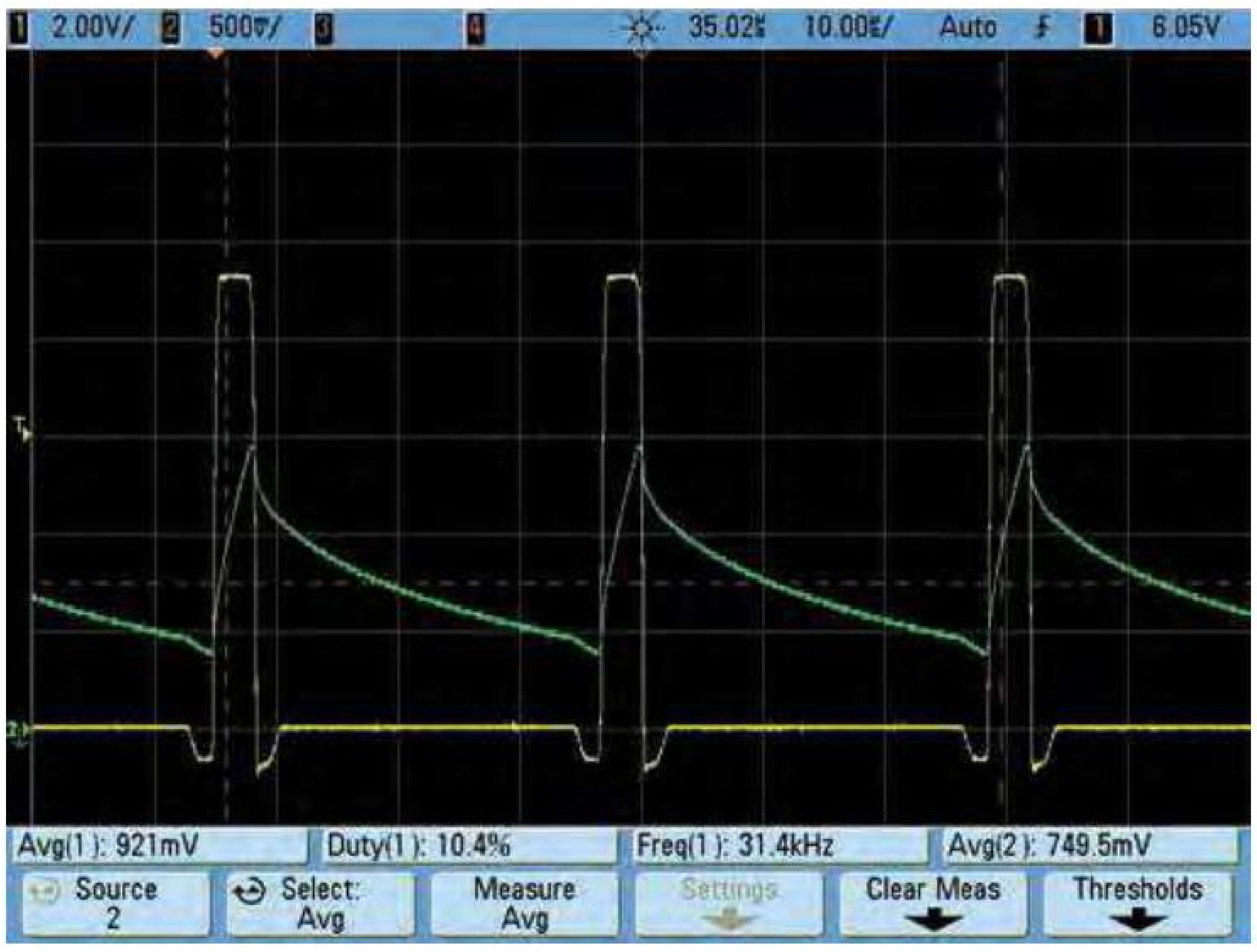
Controller power is supplied via CON1's terminals between GND and the controller supply positive input. The supply current passes through zener diode ZD1, and the input of regulator REG1 is bypassed with a 470nF capacitor.

REG1 is a low-dropout automotive 5V regulator. It is capable of withstanding a reverse polarity voltage, so it provides the circuit with reversed-supply protection. The maximum recommended operating voltage at the input of REG1 is 26V. So for use at up to 30V, ZD1 drops the voltage at the input by around 4.7V.

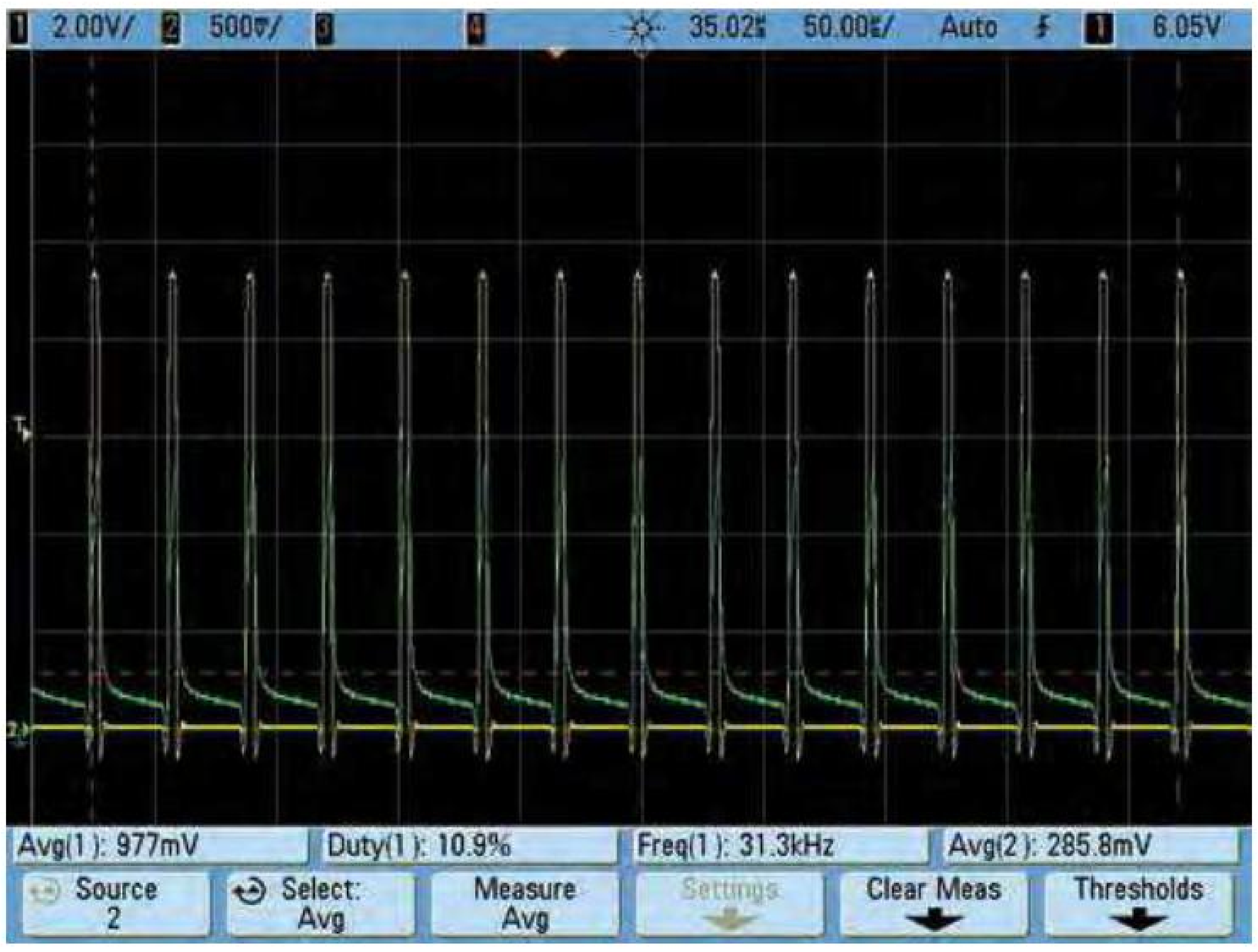
The dropout voltage for REG1 is typically 0.5V. That means it needs 5.5V at the input to ensure that the output is regulated. The addition of ZD1 means that the minimum recommended voltage for the controller is 5.5V + 4.7V = 10.2V. We round this up to 10.5V to be safe.

Note that the controller and motor positive supply connections are separate, so the motor can be run at a different voltage if required.

That means the motor supply could be outside the controller's range, and the circuit will still work as long as an appropriate controller supply voltage is applied. The two supply inputs can



Scope 5: switching back to half-bridge driving but bumping up the frequency to 31.4kHz, you can see that the average current value is hardly affected. The current level averages higher during the off-time due to the shorter off period.



Scope 6: the single-ended drive with the higher frequency suffers from the same rapid decay in current as shown in Scope 4, except this time the average current is even lower as it has less time to build during the shorter on-pulses.

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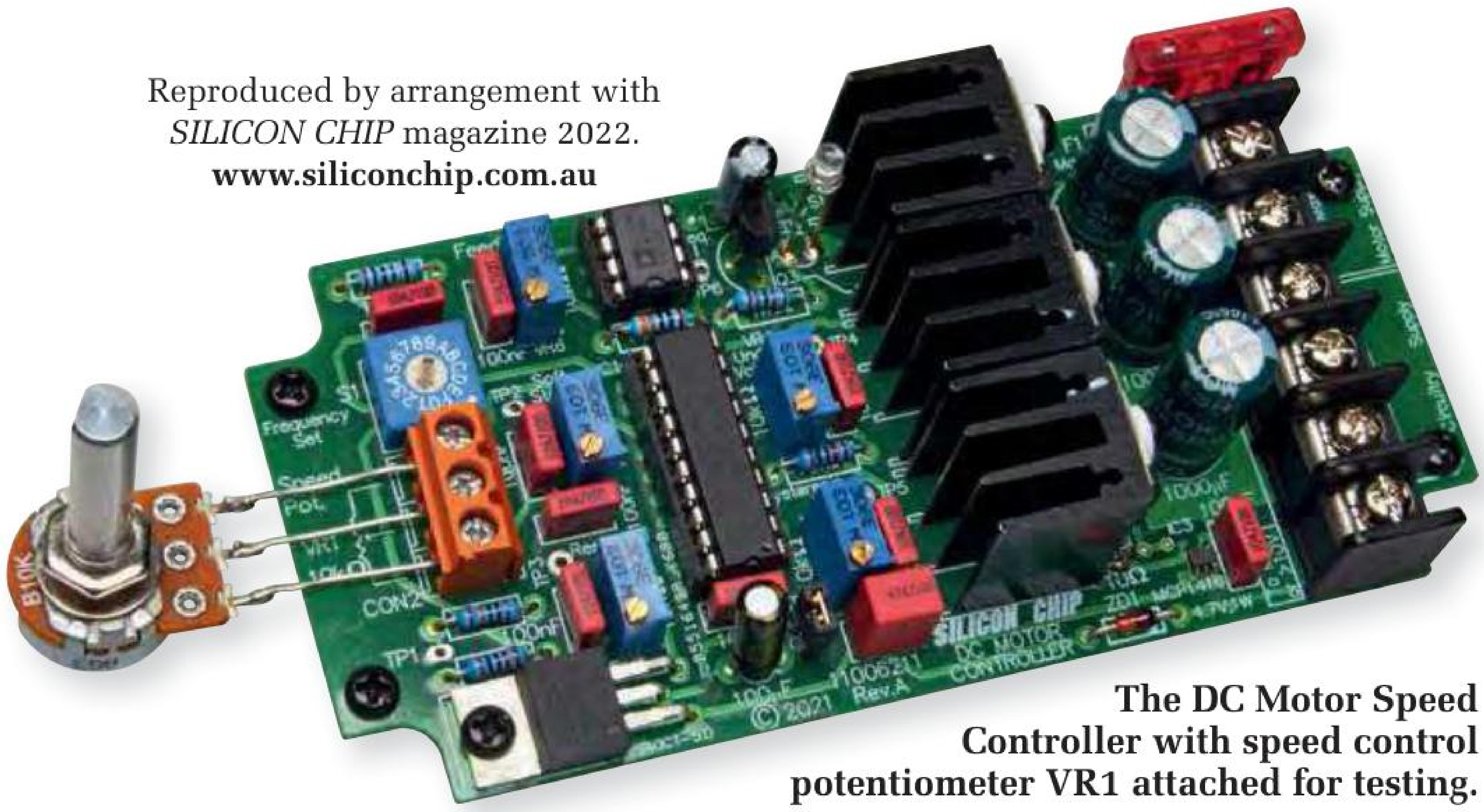


The DC Motor Speed Controller with speed control potentiometer VR1 attached for testing.

also be tied together when the motor supply voltage is within the controller's suitable range.

Construction

The *DC Motor Speed Controller* is built using a double-sided, plated-through PCB coded 11006211, measuring 122 x 58mm and available from the *PE PCB Service*. Fig.2 shows the assembly details.

Start by installing the two 10Ω surface-mount resistors and the two 0.01Ω resistors, all near Q1 and Q2. Now fit IC3, the surface-mount MOSFET gate driver. Take care when soldering this; you may need a magnifying glass and separate work light. Solder pin 1 first and check that the remaining pins are aligned correctly before soldering them.

Zener diode ZD1 can now be installed, taking care with its orientation. Follow with the seven through-hole resistors – check each one using a digital multimeter (DMM) before mounting it.

Once these parts are in place, install the socket for IC1. IC2 can be mounted using a socket, or you can solder it directly to the board. Make sure each is oriented correctly.

Now is a good time to fit MOSFET Q3, the LED and the two-way header for jumper JP1. Make sure LED1's longer lead (anode) goes into the hole at the left, marked with an 'A'. You could mount a two-pin header there instead, or solder a twin-lead cable to the board so that the LED can be chassis-mounted.

The polyester capacitors can then be inserted; it's easiest to install the electrolytic types after all the semiconductors. Follow with the trimpots, which are all multi-turn types. Orient them with the adjustment screws positioned as shown. BCD switch S1 can now be installed, with the orientation dot at the lower right.

The 3-way screw terminal block (CON2) is next on the list. Make sure it is correctly seated against the board and that its openings face outwards before soldering its pins. CON1, the 6-way screw terminal barrier block, can then go in. **Note that Altronics state these are 15A rated; however, the Dinkle data for these DT-35-B07W-XX terminals rates**

Parts List – 20A DC Motor Speed Controller

1 double-sided, plated-through PCB coded 11006211, 122 x 58mm, available from the PE PCB Service
1 UB3 Jiffy box (**optional**) **[Jaycar HB6013, HB6023, Altronics H0203]**
1 6-way 20A* PCB mount barrier screw terminals, 8.25mm pitch (CON1) **[Altronics P2106]**
1 3-way screw terminal with 5.08mm spacing (CON2)
1 10kΩ linear potentiometer (VR1)
1 knob to suit VR1
1 two-pin header, 2.54mm pitch, plus shorting block/jumper (JP1)
3 TO-220 silicone insulating washers and bushes
1 20-pin DIL IC socket for IC1
1 8-pin DIL IC socket for IC2 (**optional**)
3 TO-220 PCB-mounting heatsinks **[Jaycar HH8516, Altronics H0650]**
1 4-bit BCD switch (S1) **[Jaycar SR1220, Altronics S3001A]**
1 20A blade fuse holder (F1) **[Altronics S6040]**
1 blade fuse to suit motor (up to 20A)
4 M3 x 10mm panhead machine screws
1 M3 nut
4 6.3mm-long M3-tapped standoffs and 8 M3 x 6mm screws (**optional**; for mounting the board)
6 PC stakes (**optional**)
* Dinkle specifies these as 20A-rated; Altronics state 15A

Semiconductors

1 PIC16F1459-I/P microcontroller, DIP-20, programmed with 1100621A.hex (IC1)
1 AD627ANZ instrumentation amplifier, DIP-8 (IC2) **[element14, RS]**
1 MCP1416T-E/OT MOSFET driver, SOT-23-5 (IC3) **[RS Components 668-4216]**
1 LM2940CT-5.0 regulator, TO-220 (REG1) **[Jaycar ZV1560, Altronics Z0592]**
2 STP60NF06 N-channel MOSFETs, TO-220 (Q1,Q2) **[Jaycar ZT2450]**
1 2N7000 N-channel small signal MOSFET, TO-92 (Q3) **[Jaycar ZT2400, Altronics Z1555]**
1 3mm high-brightness LED (LED1)
1 4.7V 1W zener diode (ZD1)
1 MBR20100 dual 10A schottky diode, TO-220 (D1) **[Jaycar ZR1039]**

Capacitors

3 470µF 35V low-ESR electrolytic
1 470nF 63V MKT polyester
2 100µF 16V electrolytic
9 100nF 63V MKT polyester

Resistors (all 1/4W, 1% metal film axial unless otherwise stated)
1 100kΩ 3 10kΩ 1 2kΩ 1 1kΩ 1 330Ω
2 10Ω M3216/1206 surface mount
2 0.01Ω M6432/2512 3W surface mount **[RS Components Cat 188-0753, Vishay WFMA25120100FEA or equivalent]**
4 10kΩ top adjust multiturn trimpots (3296W style) (VR2-VR5)
1 50kΩ top adjust multiturn trimpot (3296W style) (VR6)

Table 1: PWM frequency options

BCD switch setting (S1)	PWM frequency
0	30.6Hz
1	61.3Hz
2	122.5Hz
3	245Hz
4	367.6Hz
5	490Hz
6	980Hz
7	1.96kHz
8	2.97kHz
9	3.92kHz
A	5.88kHz
B	7.84kHz
C	11.8kHz
D	15.7kHz
E	23.5kHz
F	32.4kHz

them at 20A, so they are suitable for this 20A controller.

The fuse holder is next. You can fit a monolithic holder or two separate fuse holder clips. If using individual clips, it might be a good idea to insert a fuse before soldering to ensure they are lined up correctly.

You can install PC stakes at test points TP1-TP5 and TP GND, or leave them off and probe the PCB pads directly with multimeter probes.

Installing the semiconductors

Regulator REG1 is mounted horizontally on the board. It is installed by first bending the leads to pass through their mounting holes. REG1's tab is then secured to the PCB using an M3 x 6mm machine screw and nut, after which the leads are soldered.

MOSFETs Q1 and Q2 and schottky diode D1 are mounted vertically and fastened to separate small heatsinks. The three heatsinks must be installed first, by soldering their locating pins to the relevant PCB pads. Make sure that the heatsinks are properly seated against the PCB before soldering them in place.

Then slide Q1 and Q2 into their mounting holes and, using silicone washers and insulating bushes (see Fig.3) to isolate each from the heatsink, fasten them using M3 x 10mm machine screws into the tapped holes on the heatsinks. Tighten the screws firmly, then solder their leads. Diode D1 is mounted similarly.

Now install the leftover electrolytic capacitors, taking care to orient them correctly. Finally, use your multimeter to confirm that the metal tabs of D1, Q1 and Q2 are isolated from their heatsinks.

Testing

Before inserting IC1 into its socket, check the regulator operation by applying 10.5-30V between the 0V and the controller positive supply terminals on CON1.

Measure the voltage between REG1's metal tab and its right-most lead. You should get a reading close to 5V (4.75 to 5.25V). If not, check that the input voltage at the left lead of REG1 is at least 5.5V.

If this reading is correct, switch off the power and install IC1, making sure it is oriented correctly, and none of its leads fold under the body. If you used a socket for IC2, plug it in now.

At this stage, it is a good idea to wire potentiometer VR1 to CON2. You will also need to insert the fuse to continue testing. The fuse should be rated to suit the motor; if it is a 1A-rated motor, install a 1A fuse; for a 20A motor, use a 20A fuse... and so on.

Next, wind the curve adjustment trimpot VR3 fully anti-clockwise. You can find this position by winding at least 20 turns anti-clockwise or until a faint clicking sound is heard. When the circuit is powered, the voltage reading between TP3 and GND should be very close to 0V.

Low-voltage cut-out testing

When power is applied, the LED will flash at 1Hz because there is no power connected to the motor supply.

Trimpot VR4 sets the low-voltage cut-out. With a multimeter connected between TP4 and TP GND, adjust VR4 for one-tenth of the desired low cut-out voltage. So for a low voltage cut-out at 11.5V (a safe level for most 12V lead-acid batteries), adjust TP4 until you get a reading of 1.15V.

Adjusting the hysteresis is similar, using trimpot VR5 and measuring at TP5. The hysteresis is the voltage measured at TP5 (not 1/10th as before). So for a 1V hysteresis, set TP5 to 1V. Hysteresis can be set for up to 5V, but 1V is a reasonable starting point. With the recommended 11.5V cut-out voltage, that means the battery voltage needs to rise above 12.5V (about half-charge) before operation resumes.

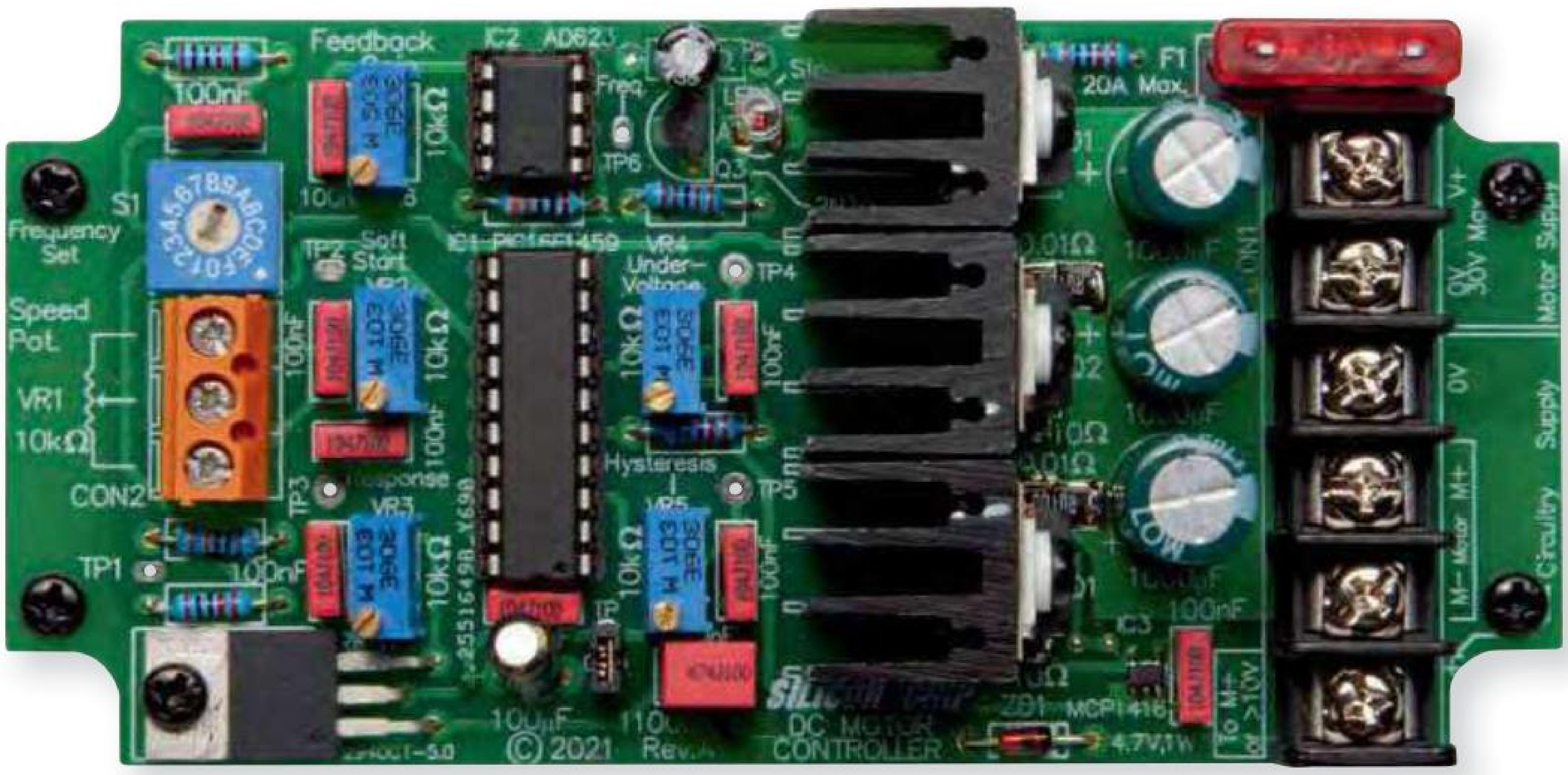
If you have an adjustable power supply, the low-battery cut-out can be tested. Connect this supply between the motor supply positive and 0V, and rotate VR1 fully clockwise. The LED will light up when the supply voltage is in the operating range and flash when a low voltage is detected.

Set the supply to more than the low voltage cut-out setting plus the hysteresis setting, so the low-voltage cut-out will not initially activate. Then reduce the voltage to the cut-out voltage. Note that the low-voltage protection will take about 10s to occur once the supply is below the threshold. LED1 should then flash at 1Hz.

Slowly increase the supply to just over the threshold plus the hysteresis setting value (12.5V in our example), and LED1 should light fully. If necessary, adjust VR4 and VR5 to get it to cut out and in at precisely the voltages you require.

Soft-start setting

Adjust VR2 for the required soft-start rate. Typically, 5V at TP2 is suitable, giving a maximum two-second soft-start period. You can reduce this for faster starting, or disable soft starting with 0V measured at TP2.



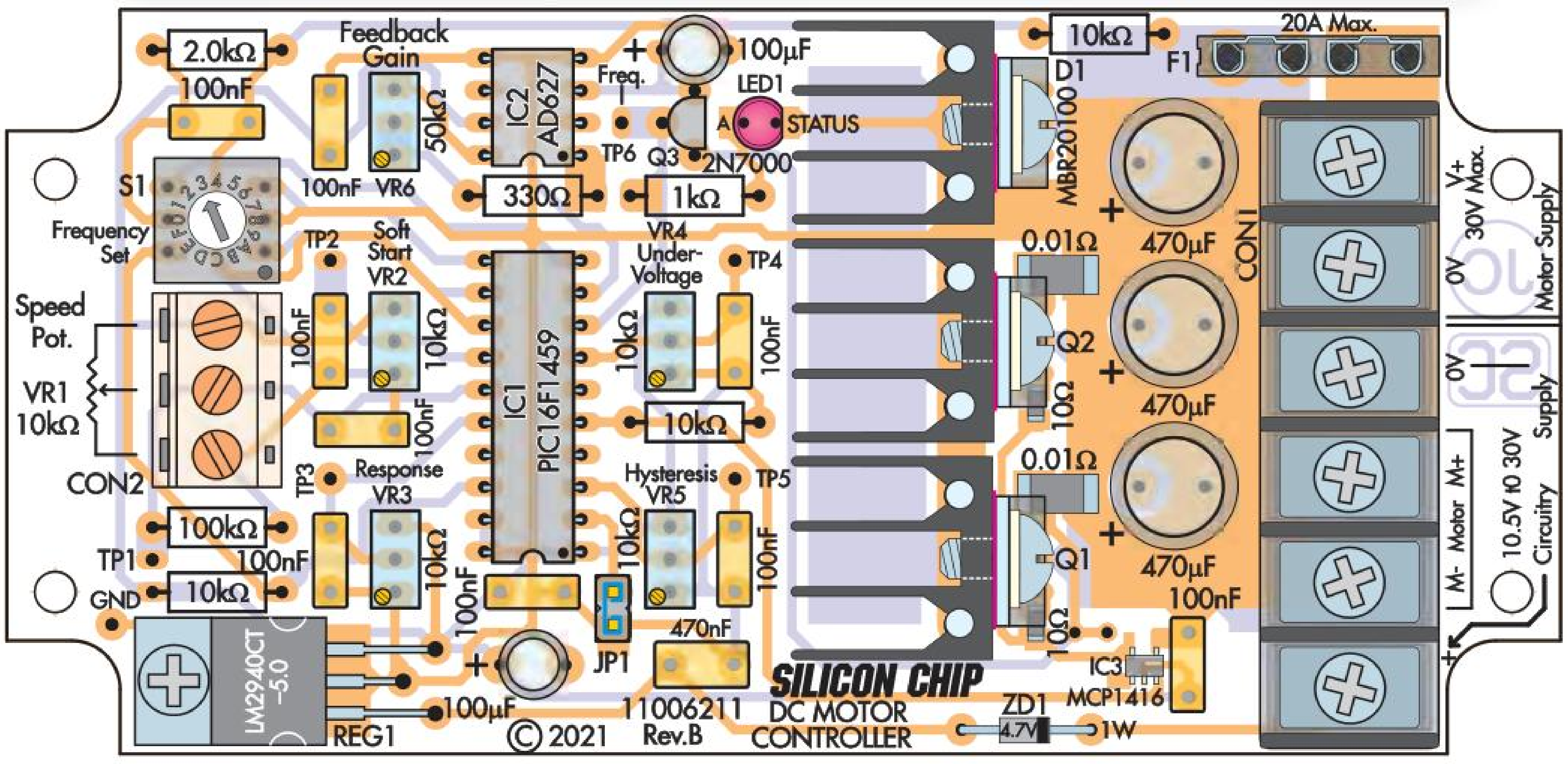


Fig.2: the Speed Controller PCB is relatively compact and uses just five SMD parts: four resistors and MOSFET driver IC3. MOSFETs Q1 and Q2 and diode D1 attach to PCB-mounting heatsinks for cooling. During assembly, watch the polarity of the three ICs, diode ZD1, the electrolytic capacitors and BCD switch S1.

The disadvantage of back-EMF-based speed feedback

Typically, a DC motor acts as a generator when the power is switched off. When using PWM drive, this generated voltage or back EMF (Electromotive Force) occurs repetitively when the driving MOSFETs are switched off. But the induced voltage is not developed immediately after switch-off; it does not happen until the stored charge in the inductance of the motor windings dissipates.

In many speed controllers, the back EMF voltage is used to stabilise the speed with varying load. As the motor is loaded, the speed and back EMF reduce, and this change is used to provide feedback that increases the PWM duty cycle to maintain speed under load.

However, with higher PWM frequencies, the back EMF voltage appears much later in the PWM cycle; sometimes, it is not developed until after the MOSFETs are switched on again, so it is impossible to sense the back EMF.

Compare scope grabs Scope 7 and Scope 8. They are the same except that the PWM frequency is just under 3kHz in Scope 7 and nearly 12kHz in Scope 8. You can see the back EMF 'shelf' appear about 80µs after switch-off in Scope 7, but it is barely visible in Scope 8 and would not be present at all with a higher switching frequency.

The lack of back-EMF at high PWM frequencies means that we need to use a different way of detecting motor load. The easiest alternative is to measure the motor current. We only do this while the motor is driven by amplifying the voltage across a low-value shunt resistor in series with the motor.

Using feedback control based on measuring current, the PWM duty cycle can be increased whenever the motor is loaded. This tends to overcome the shortcomings of low torque at high frequencies and lower duty cycles, to some extent at least.

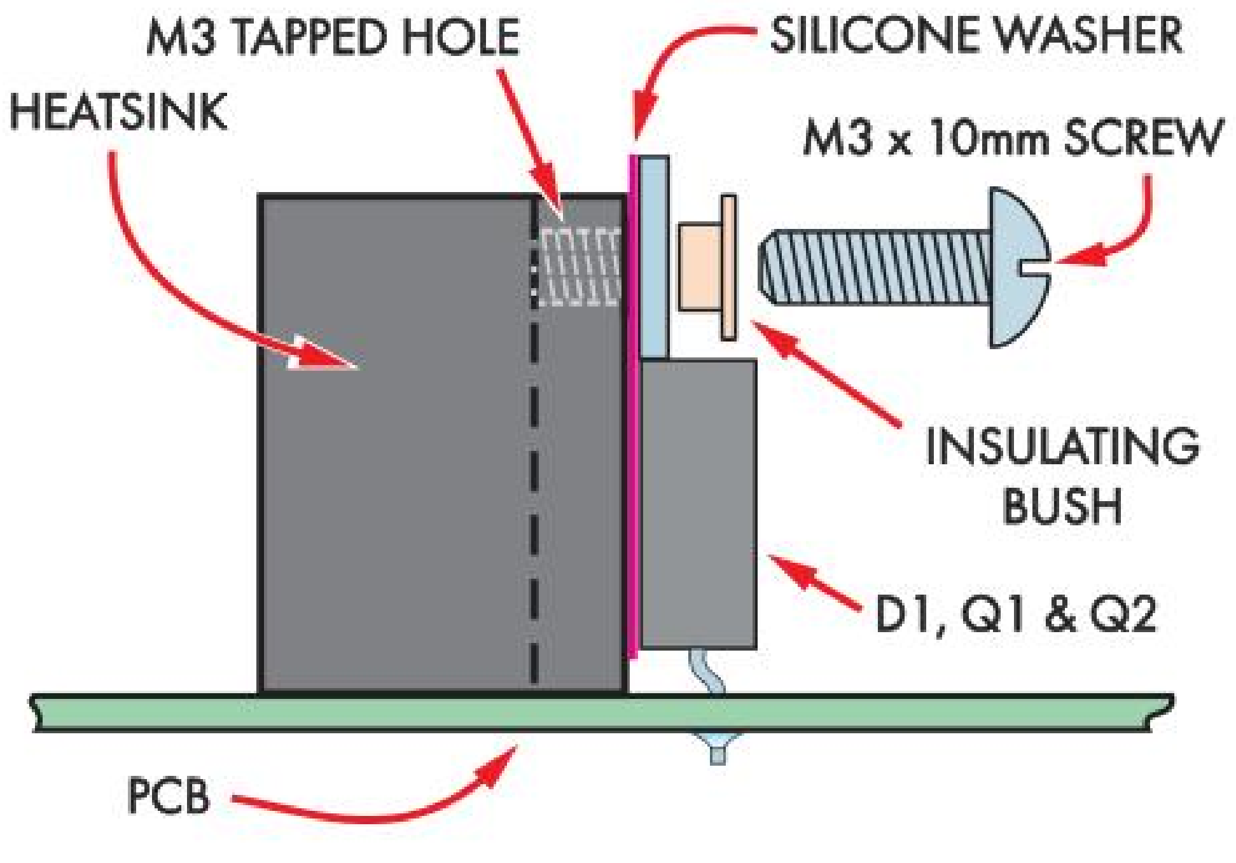


Fig.3: this side view shows how the TO-220 package devices are mounted to the heatsinks. The hole in the heatsink is pre-tapped. The heatsinks are connected to ground via the PCB and mounting pins, so you do need the insulating washers and bushes.

Curve adjustments

VR3 sets the curve adjustment. This is off when VR3 is wound fully anti-clockwise, with 0V at TP3. Rotating VR3 clockwise will increase the curve adjustment. For settings above 2.5V, see the optional motor disconnection detection section below.

As mentioned earlier, the curve setting provides high-frequency operation improvements when JP1 is out of circuit or low-frequency operation improvements with JP1 inserted.

With JP1 out, VR3 increases the minimum duty cycle for low settings of VR1. To make the adjustment, rotate speed potentiometer VR1 slightly clockwise from fully anticlockwise, giving a reading of just over 20mV at TP1. Then adjust VR3 clockwise until the motor just starts to run.

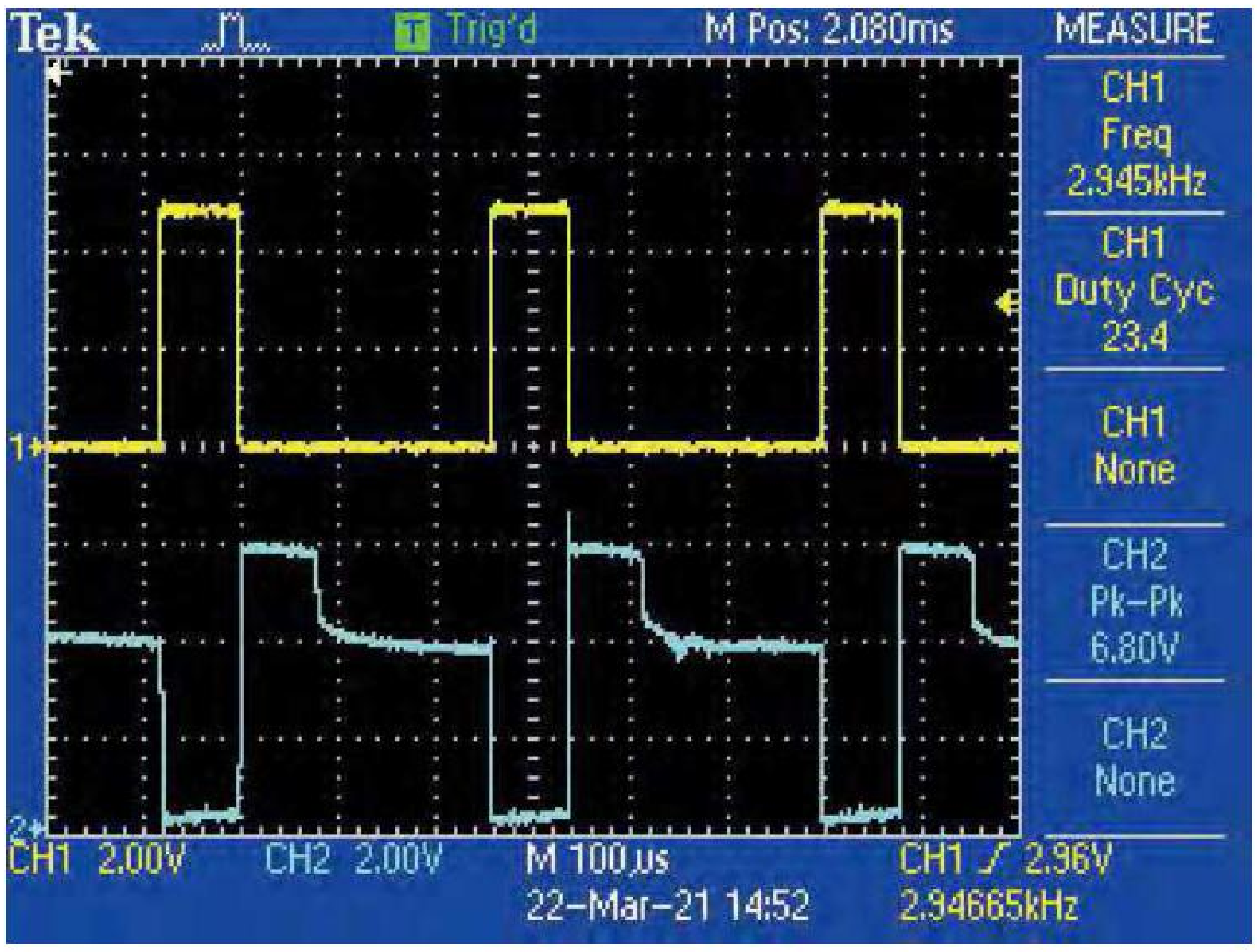
Adjust the gain control (VR6) for best motor control for maintaining motor speed under load. Clockwise will give more gain, and anti-clockwise will set a lower gain. Setting the gain too high can cause the motor speed to become unstable.

Set the PWM frequency to a value that you find best for the motor. This will be a compromise between motor control performance and the amount of PWM noise made by the motor. Very low frequencies can cause the motor to run coarsely. Very high frequencies will improve smoothness, but can reduce torque at lower settings unless the feedback control is adjusted to give better performance under load.

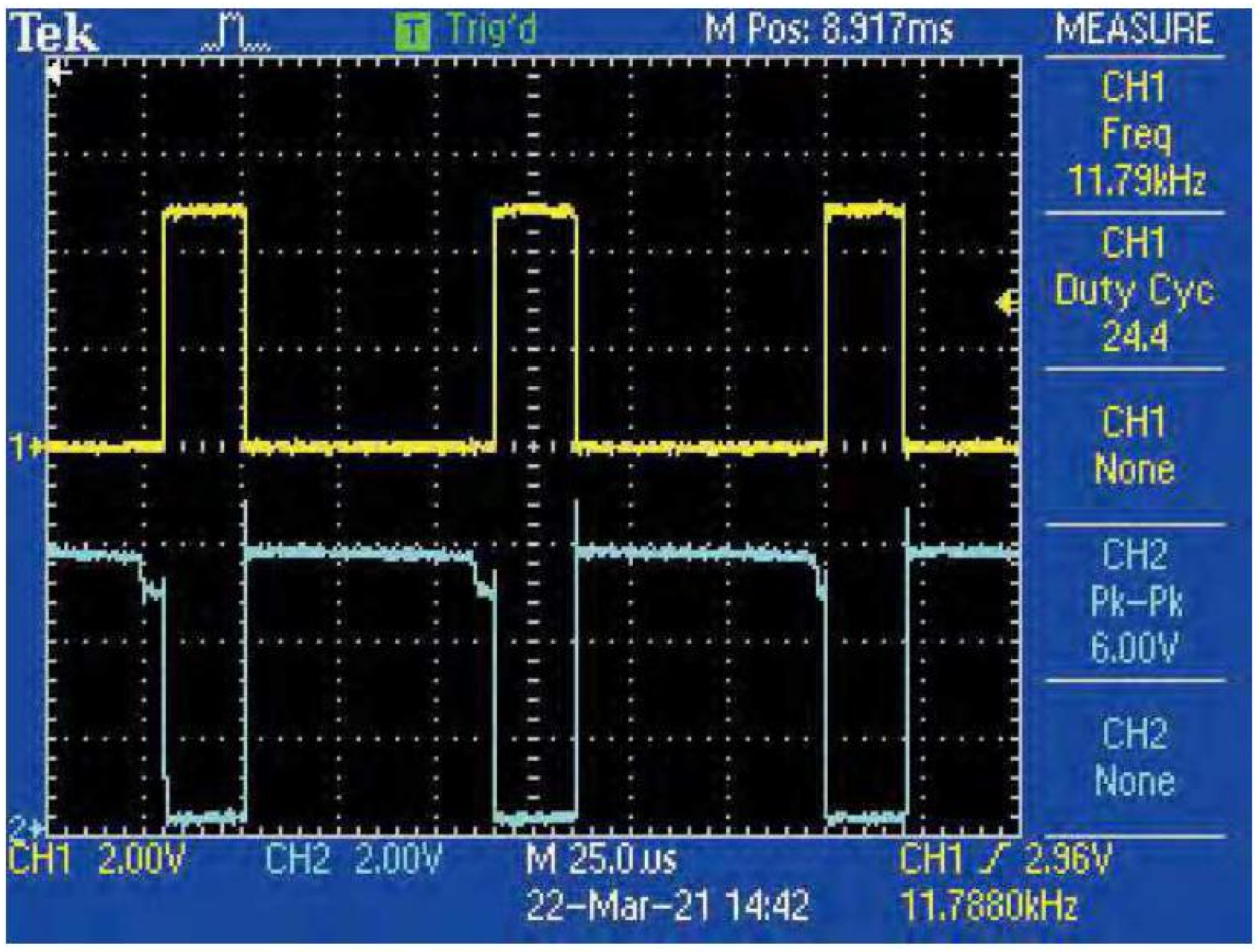
Adjust the response trim pot (VR3) to give the best speed control range for VR1. When the PWM frequency is low, you might find that the motor speed can increase sharply when winding VR1 up from zero, especially when there is high feedback gain. Adjusting the response using VR3 with JP1 inserted can reduce this snap-on effect. Start from 0V (at TP3) and adjust VR3 until the motor runs well at low duty cycles, without the snap-on effect.

Motor disconnection detection

If you want this option, the curve adjustment trimpot (VR3) is set in the opposite manner. There is no curve adjustment when VR3 is fully clockwise (5V at TP3), and the curve adjustment increases as VR3 is wound further anti-clockwise. It is usable down to 2.5V at TP3.



Scope 7: with a PWM frequency just under 3kHz, there is sufficient time for back-EMF sensing. The motor voltage shoots up immediately after the MOSFETs switch off, then falls back to a lower plateau once the magnetic field has decayed and back-EMF starts to become dominant.



Scope 8: with a PWM frequency of nearly 12kHz – the back-EMF voltage is barely visible just before the start of the next pulse. It would be impractical to sample the back-EMF voltage at this frequency for this motor, and impossible at higher frequencies.

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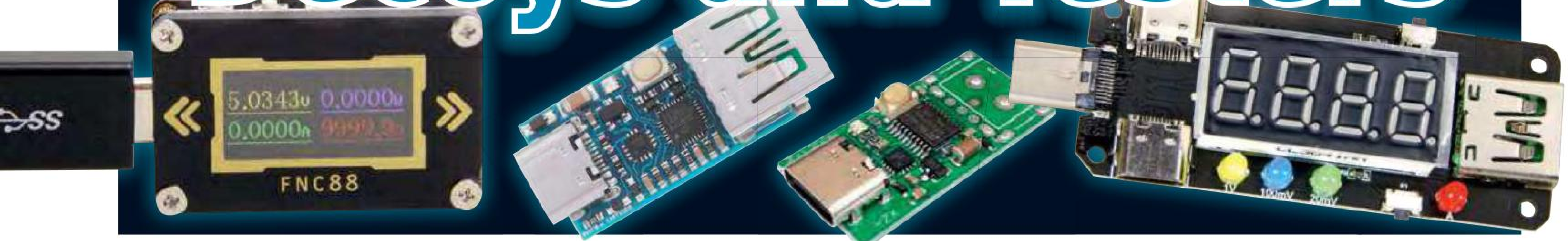
Using Cheap Asian Electronic Modules

By Jim Rowe

USB-PD Triggers, Decoys and Testers



Left-to-right: the FNC88, WITRN UPD005, ZY12PDN-1 and XY-WPDT

Following on from the article last month on USB-PD charging modules, we shall now examine some of the many low-cost PD trigger/decoy and tester modules that have appeared recently. These allow you to take full advantage of the USB-PD chargers' capabilities and use them as versatile and efficient power supplies.

A **PD trigger is an interface** circuit that can manage the negotiating protocols necessary to request the required voltage and current levels from a USB-PD supply, as described last month. As soon as it is connected to a USB-PD compatible host, it engages with the host via the CC1 or CC2 channels to achieve the voltage and current levels that are needed – assuming these are available.

Some of these modules are also known as 'decoy' or 'poll detectors'. These terms all seem to mean much the same thing as PD trigger.

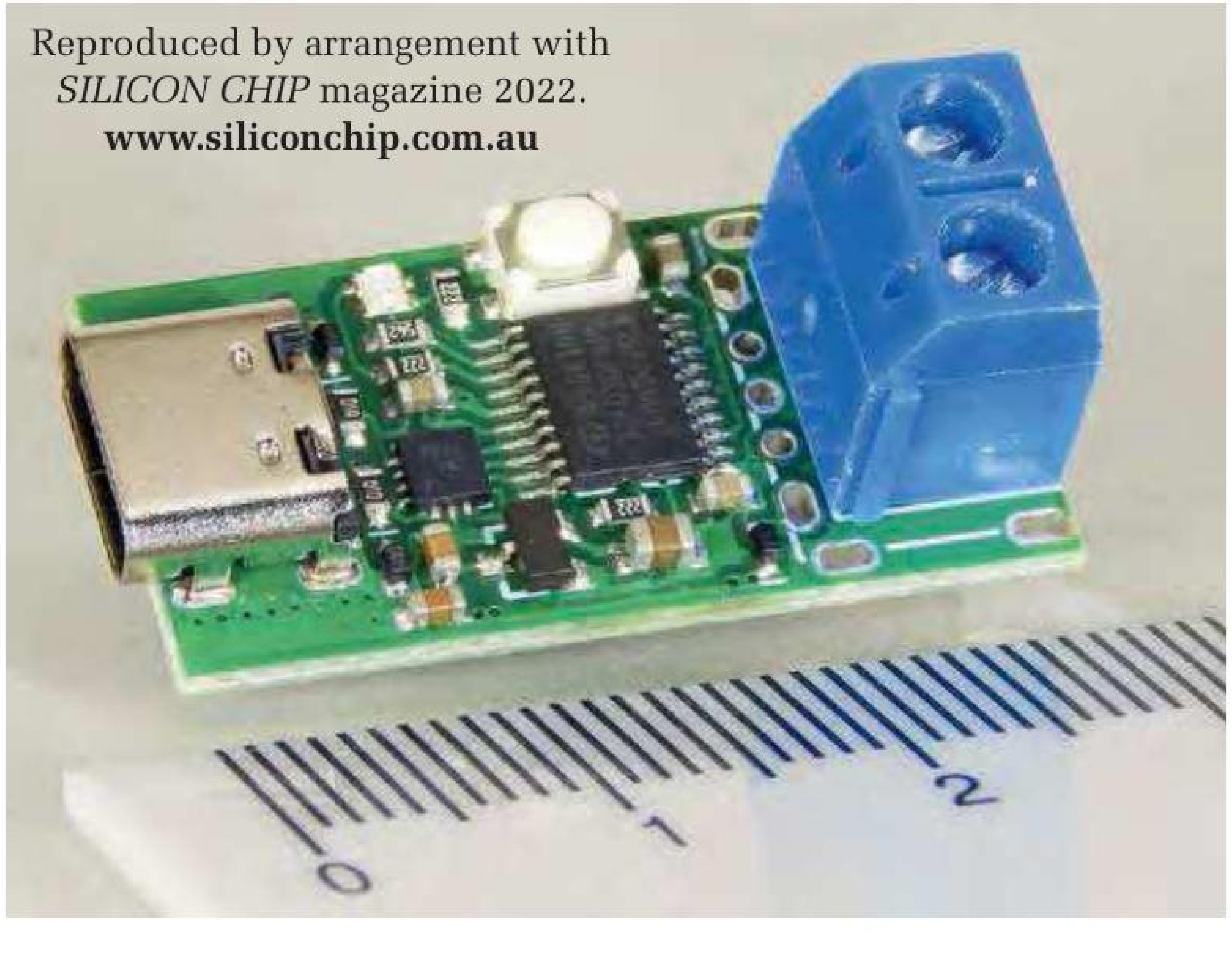
Another variant combines the functions of a trigger/decoy module with those of a USB-PD digital meter, so it can display the available or selected voltage(s) and current(s).

We'll start by looking at the smallest, simplest and cheapest of the trigger modules.

ZY12PDN-3 'naked' PD trigger

The ZY12PDN-3 module is tiny – see the photos. Everything is mounted on a PCB measuring just 30 x 15mm, with the USB-C input socket at one end and a small two-way screw terminal block at the other end as the power output.

The ZY12PDN trigger module is available in three versions, which differ only in their output connector or lack thereof. Instead of the screw terminal block of the ZY12PDN-3, the ZY12PDN-2 has a USB Type-A socket, while the ZY12PDN-1 has no output connector at all.



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The ZY12PDN-3 PD trigger, shown enlarged for clarity. There are two other versions of this module with either a USB Type-A socket or no connector fitted instead of the screw terminal block.

The trigger circuitry on the PCB uses two main chips: an STM32F030F4P6 microcontroller and a PBAFH device, which is likely the USB physical layer interface.

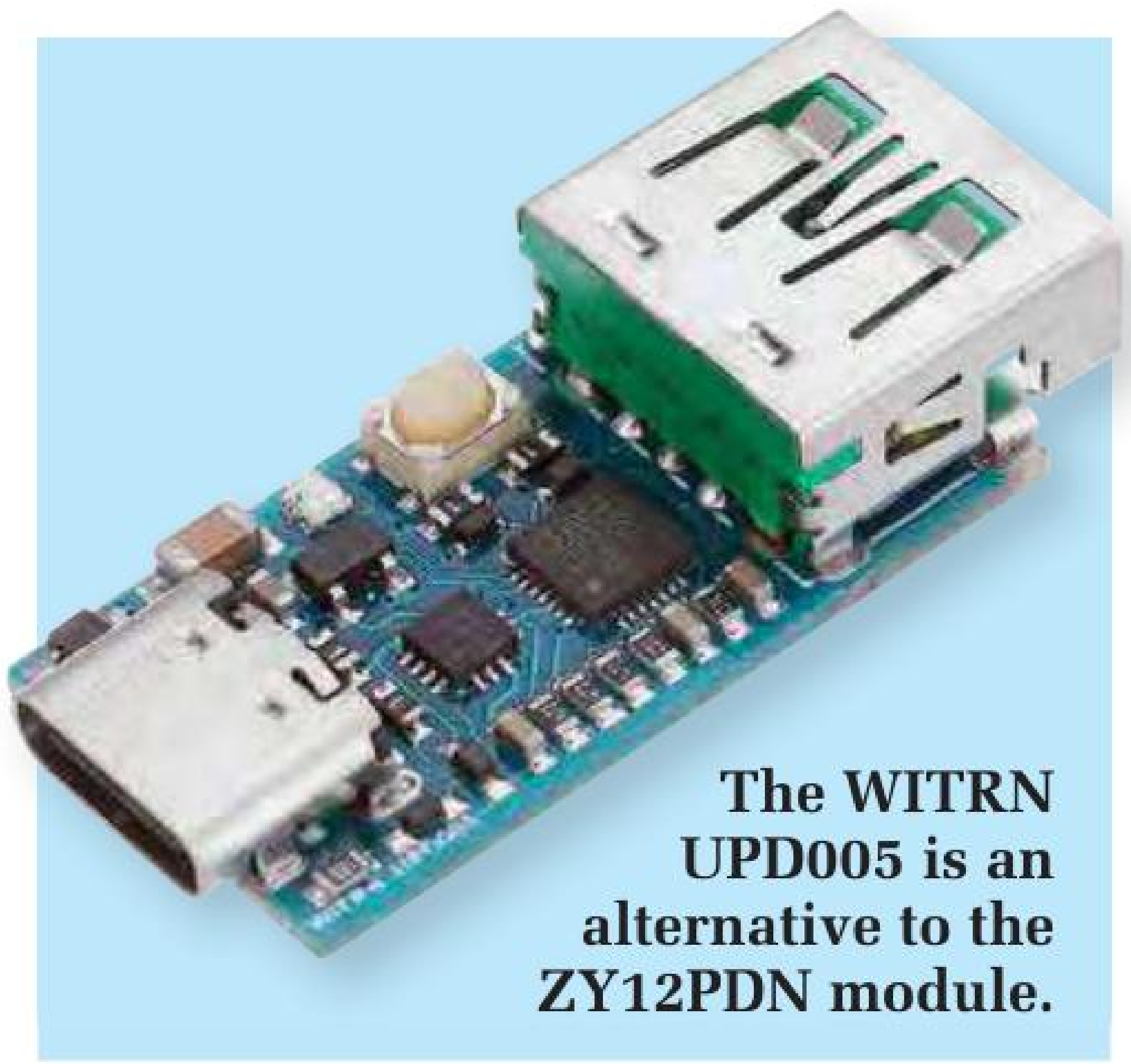
There's also a small pushbutton switch that can be used to select the voltage and power level required from the USB-C PD host, and an RGB LED to indicate the selected voltage/power level.

When the ZY12PDN is first connected to the PD host, the LED glows red to indicate the default 5V supply voltage. If you then press the button, it will attempt to select a 9V supply. If the PD host has this voltage available, it will switch its output to 9V, and the LED will change to yellow. If you press the pushbutton again, this will attempt to change the supply voltage to 12V. If the PD host has this voltage available, it will switch its output to 12V, and the LED will change to green. Further button presses will change the voltage to 15V (light blue), then 20V (dark blue) – again, assuming the host can supply these voltages.

If the host doesn't have one of the voltages that is requested, then the LED will glow purple, and the voltage will stay at the highest voltage which is available.

There's also a 'demo' mode, where the LED glows white and the supply voltage cycles through the available levels at approximately 1Hz.

If you plug the ZY12PDN into a host port that does not support USB-PD voltage and power negotiation, after



The WITRN UPD005 is an alternative to the ZY12PDN module.

about four seconds, the LED will flash blue to warn you that there is no USB-PD support. However, it will still pass through the normal 5V V_{BUS} power.

So, the ZY12PDN trigger module essentially provides the ability to manually select the voltage from a USB-PD power source. And it does this for a cost of around £5-£8, depending on how many you order and from which Internet supplier.

It's a bit limited in terms of the voltages you can request and has no provision for taking advantage of PPS 'fine tuning', but if you just need the ability to manually select one of the main PD voltage levels, it is a good choice. For example, you could use it in combination with a computer or USB charger as a very basic bench supply to power something like a breadboard.

There are other 'naked' trigger modules available that are very similar to the ZY12PDN. One example is the WITRN UPD005 V20, available from suppliers like Banggood for much the same price as the ZY12PDN. I obtained one of these and tried it out, and it did the job just as well as the ZY12PDN.

FNC88 PD trigger module and meter

The FNC88 PD trigger is slightly larger than the ZY12PDN, but not by much, especially considering that it also includes a digital meter. It measures just 40 x 25 x 10mm, not including the USB-C input plug. And it's not exactly 'naked' either, with a shield PCB mounted 3mm below the main PCB and a protective plastic sheet above the 24mm diagonal colour LCD screen.

It's made by FNIRSI Technology in Shenzhen, China, and is available from quite a few internet suppliers, including Banggood, for around £20 plus delivery.

It comes in a protective case with a clear window, measuring 90 x 62 x 18mm. This should make it sturdy enough for portable use.

The FNC88 PD unit has USB-PD bidirectional capability, and this also applies to the built-in digital meter. So PD triggering and measurements can be made using either the USB-C plug at one end or the USB-C socket at the other end as the power source, with the opposite end connected to the 'sink' device.

This also means that the FNC88 can be connected between a USB-C cable and the USB-C connector of either a host or sink device.

On one side of the FNC88's main PCB, there's a mini USB-C socket, which extends its capabilities to measuring the current drawn by devices with that type of USB connector. Then on the other ('top') side are three tiny pushbutton switches, and an equally small slider switch.

The slider switch is to enable or disable the trigger's USB-PD protocol communicating ability, while the three pushbuttons are used to select the functions of the digital meter.

The specified supply voltage range of the FNC88 is 4-24V, and its current range extends from 0 to 5A, so it's capable of dealing with all devices conforming to the current USB-PD specification.

The voltage measurement resolution and accuracy are specified as 0.1mV and ±(0.5% + 2LSD), while the current measurement resolution and accuracy are specified as 0.1mA and ±(1% + 2LSD). Quite impressive!

Other features of the FNC88 include measurement and display of:

- The power being drawn by the sink device (0-150W, with a resolution of 10mW).
- The charge delivered to a battery over a known charging time (0-99,999.9mAh with a resolution of 0.1mAh).
- The energy delivered to a battery or other sink device over a known delivery time (0-9999.999Wh with a resolution of 1mWh).
- The ability to record measurements over a period of 0-999 hours, 59 minutes and 59 seconds with a resolution of one second and an accuracy of 10 seconds/hour.

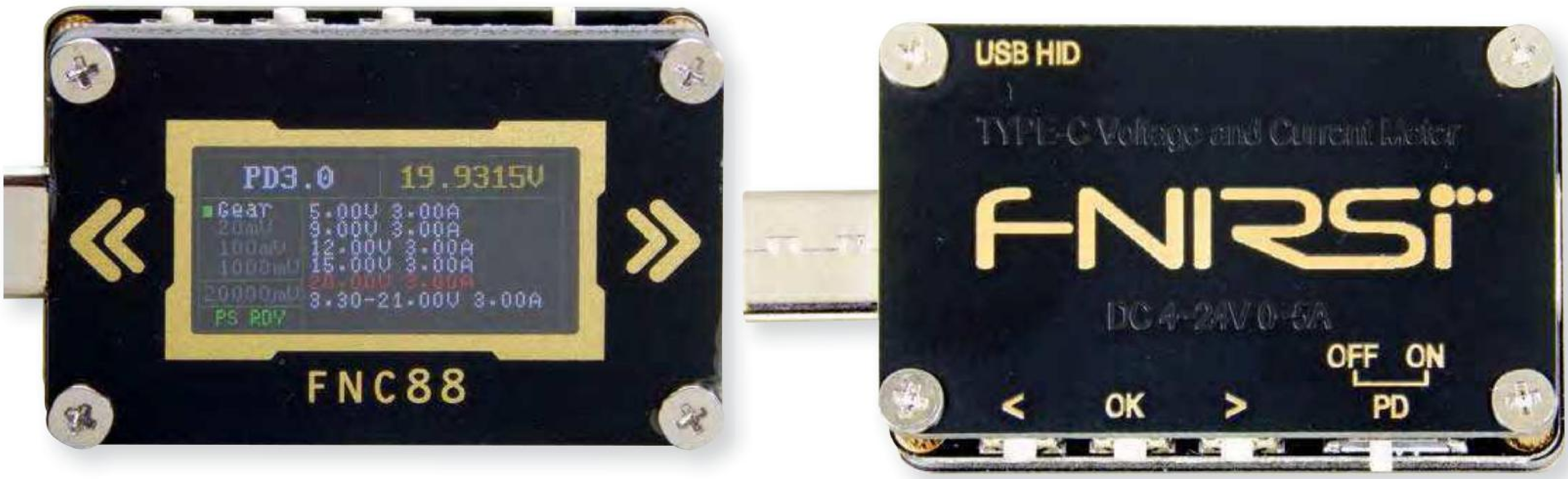
The PD trigger section of the FNC88 supports these protocols: QC2.0, QC3.0, FCP, SCP, AFC and PD 3.0.

Although the FNC88 does not come with any user operating manual, you can download one as a PDF file from the FNIRSI website at: **https://bit.ly/pe-sep22-fnc99**

I gave the FNC88 a quick rundown, comparing its voltage and current readings with those from my Agilent U1251B DMM. This showed that the accuracy and resolution of the FNC88's digital meter were within their claimed figures.

So overall, the FNC88 seems to be a very capable and useful device. My only real complaint is that you need either good eyesight or a strong magnifying glass to read the display on its 24mm diagonal LCD screen.

There are several other USB-C PD trigger/DM devices available from Banggood and other internet vendors. A good example is the Riden TC66, which is almost precisely the same size as the FNC88 and very similar in its functions and facilities. It's available for much the same cost as the FNC88.



The front and rear view of the FNC88 module; you can just see the three function buttons at the bottom of the rear view, along with the switch that connects the built-in PD chip to the CC1 pin.

This side of the FNC88 module has a micro USB interface which is only used to flash the firmware.

I have seen a suggestion on the internet that the FNC88 might be a knock-off of the TC66, or vice versa. See: **https://bit.ly/pe-sep22-tc66**

XY-WPDT trigger unit and meter

Another USB-PD trigger/meter unit available from many internet suppliers, including Banggood, is the XY-WPDT. At the time of writing, Banggood was selling it as a kit for only £7.50 including delivery. It is made by the same firm in China which makes the XY-PDS100 'quick charger' we looked at last month.

Although the XY-WPDT comes as a kit, assembling it is not difficult and doesn't involve any soldering – just the use of a very small Philips-head screwdriver, which is included in the kit. The PCB itself is already assembled, so all that remains is fitting the front and rear panels around it using the M2.5 screws and tapped spacers provided.

When you complete the assembly, the XY-WPDT measures a modest 61 x 25 x 11mm (not including the USB-C plug protruding from its input end). It's only a little larger than the FNC88.

The main output is via a USB Type-A socket at the opposite end of the unit to the USB-C input plug, and the XY-WPDT comes with a 100mm-long output cable with a Type-A plug at one end and a 2.5mm inner diameter concentric power connector at the other.

There are also a couple of USB-C sockets on the unit itself near the USB-C input plug, one on each side, arranged so that the XY-WPDT can measure the voltage and current passing between them.

The specifications of the XY-WPDT are not all that different from that of the FNC88. It can negotiate an output voltage between 4V and 20V using either PD 2.0 or PD 3.0 protocols. It can adjust the voltage in either 1V, 100mV or 20mV increments or decrements if the PD host can respond to PPS negotiation (like the XY-PDS100).

The meter function can measure the voltage with a rated resolution of 10mV and a precision of 0.3%, and current with a rated resolution of 1mA and a precision of 0.5%. Not quite as good as the FNC88, but still very useful.

The readout of the XY-WPDT is a 4-digit 7-segment LED display with 9mm-high digits, so although it is not as fancy as that of the FNC88, it's significantly easier to read.

Function switching is done via two tiny pushbutton switches (K1 and K2), one on either side of the unit. There are also four indicator LEDs; three indicate the voltage steps in PPS mode (1V/100mV/20mV), with the remaining one indicating current measurement mode.

Like the FNC88, the XY-WPDT does not come with any operating manual, nor could I find a manual on the internet. The only information on using it seemed to be in the XY-WPDT follow-up info on the Banggood website, which turned out to be rather terse and not easy to follow.

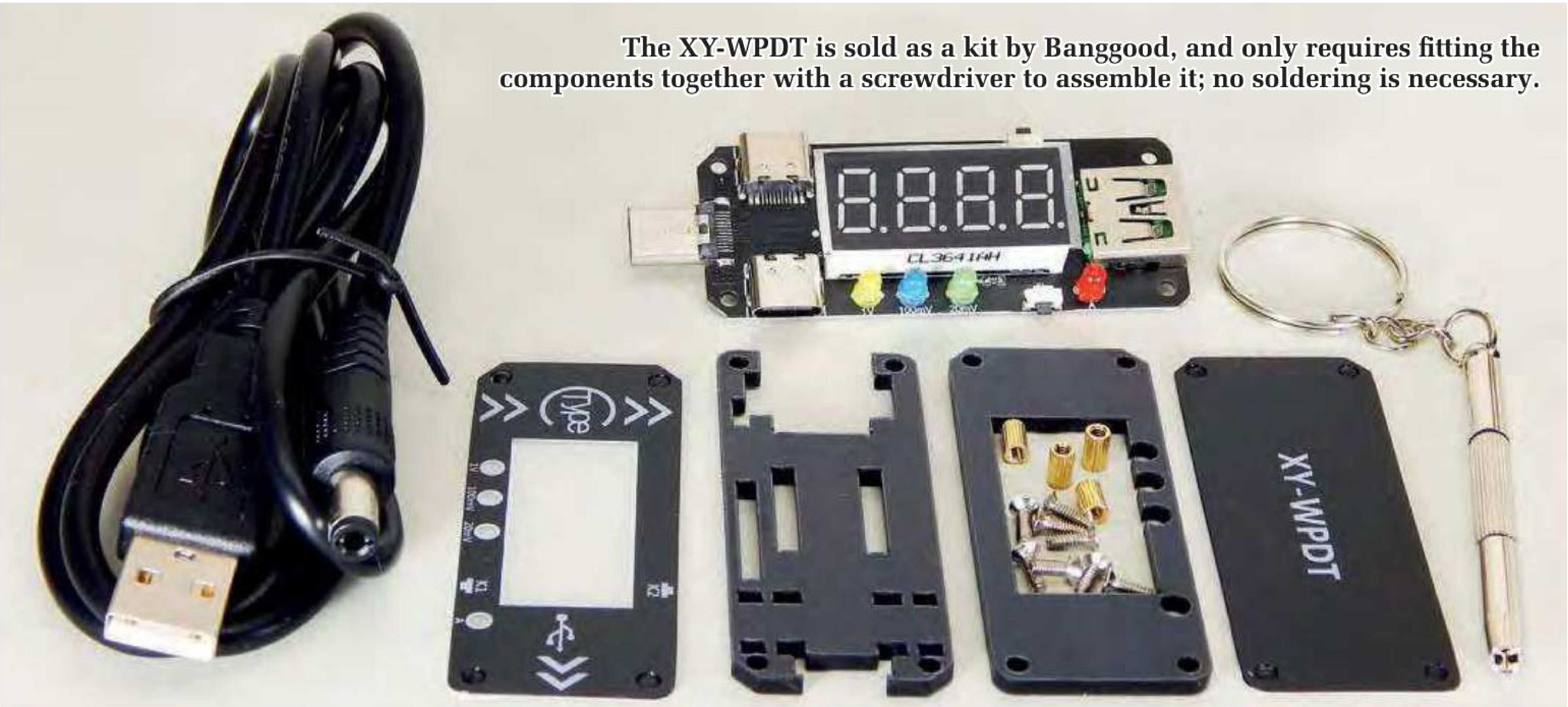
I gave the XY-WPDT a quick checkout coupled to the XY-PDS100 PD charger, and the results were very close to the rated figures for resolution and precision of both voltage and current.

Overall then, the XY-WPDT PD trigger/meter is quite a good performer, and very good value for money. My only real complaint is that the method it uses to select the voltage mode using the two tiny pushbuttons K1 and K2 is really tricky, with various short and long presses on each button making it not at all easy to set the XY-WPDT to a particular voltage level, especially in PPS mode.

This seems to be because both buttons have different functions according to how long they're pressed, so you can easily flip things into a different mode without meaning to.

In theory, the combination of the XY-WPDT and the XY-PDS100 should make a digitally adjustable DC power supply with its output variable to any voltage between 4V and 20V, but this isn't all that easy in practice.

It would be a lot easier if the two tiny pushbuttons were increased in



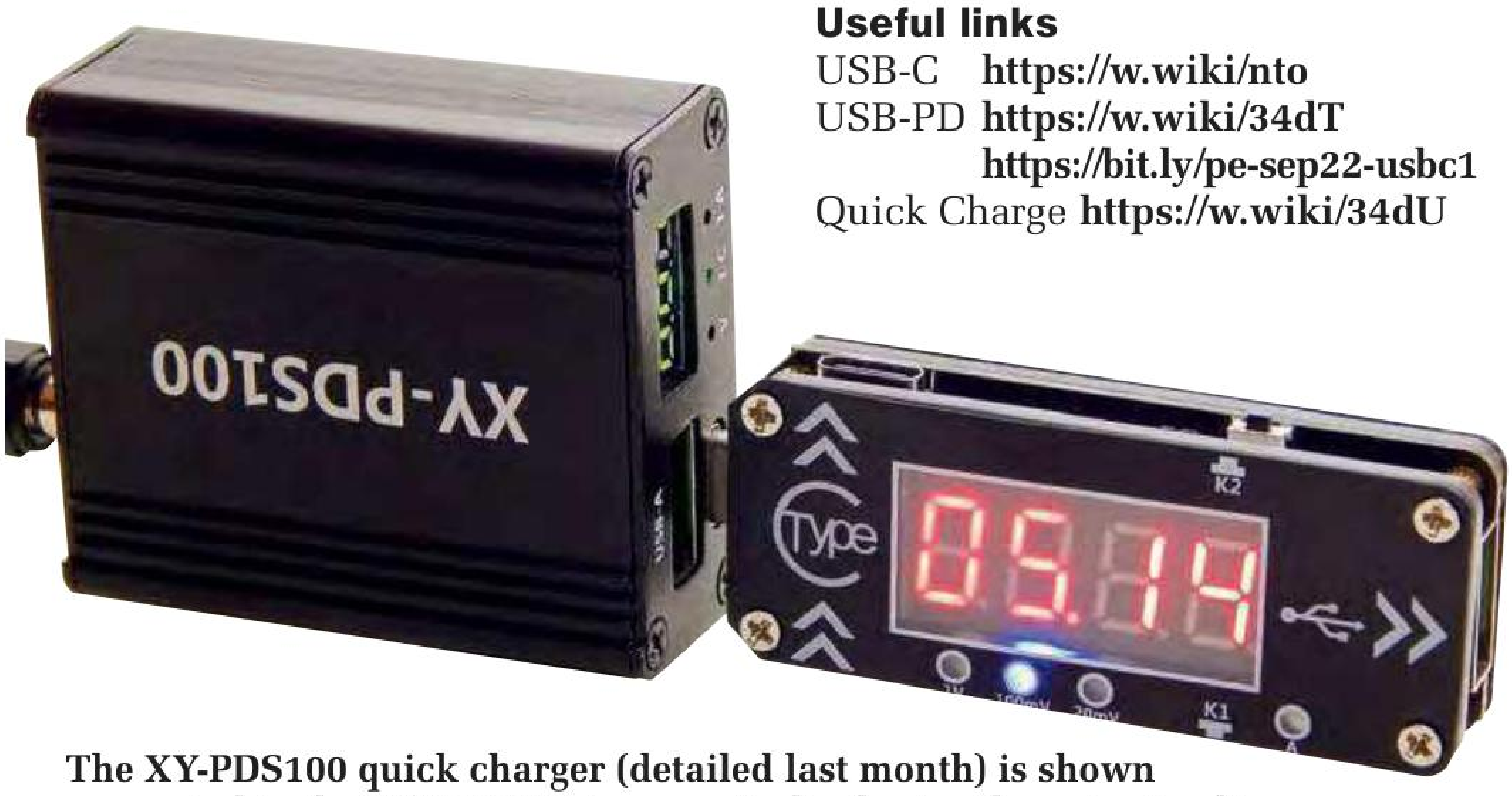
The XY-WPDT is sold as a kit by Banggood, and only requires fitting the components together with a screwdriver to assemble it; no soldering is necessary.

number, with a smaller number of functions per individual button and less dependence on the time they are pressed.

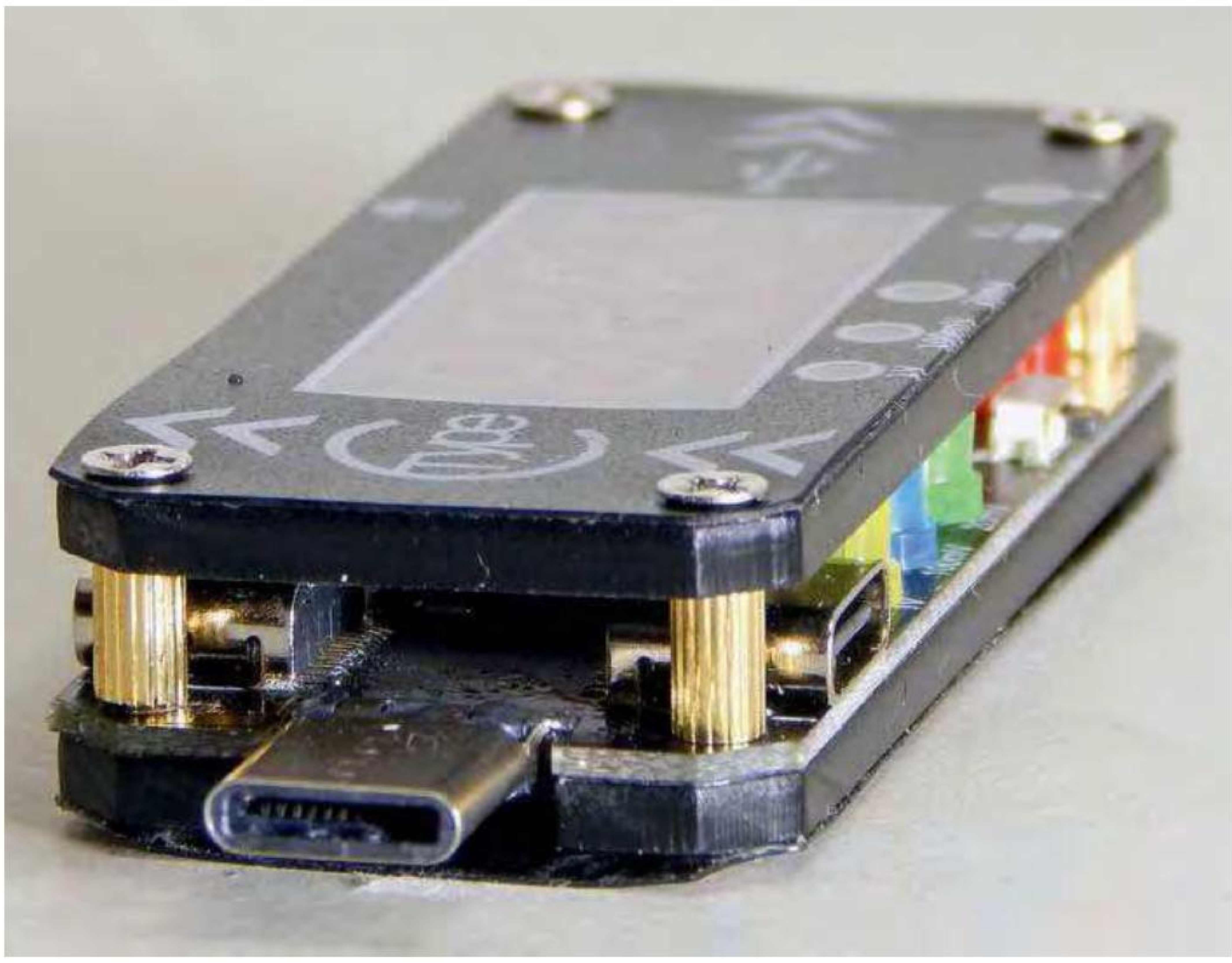
But for applications where you want to 'set and forget', it works acceptably well and provides excellent value for money.

Useful links

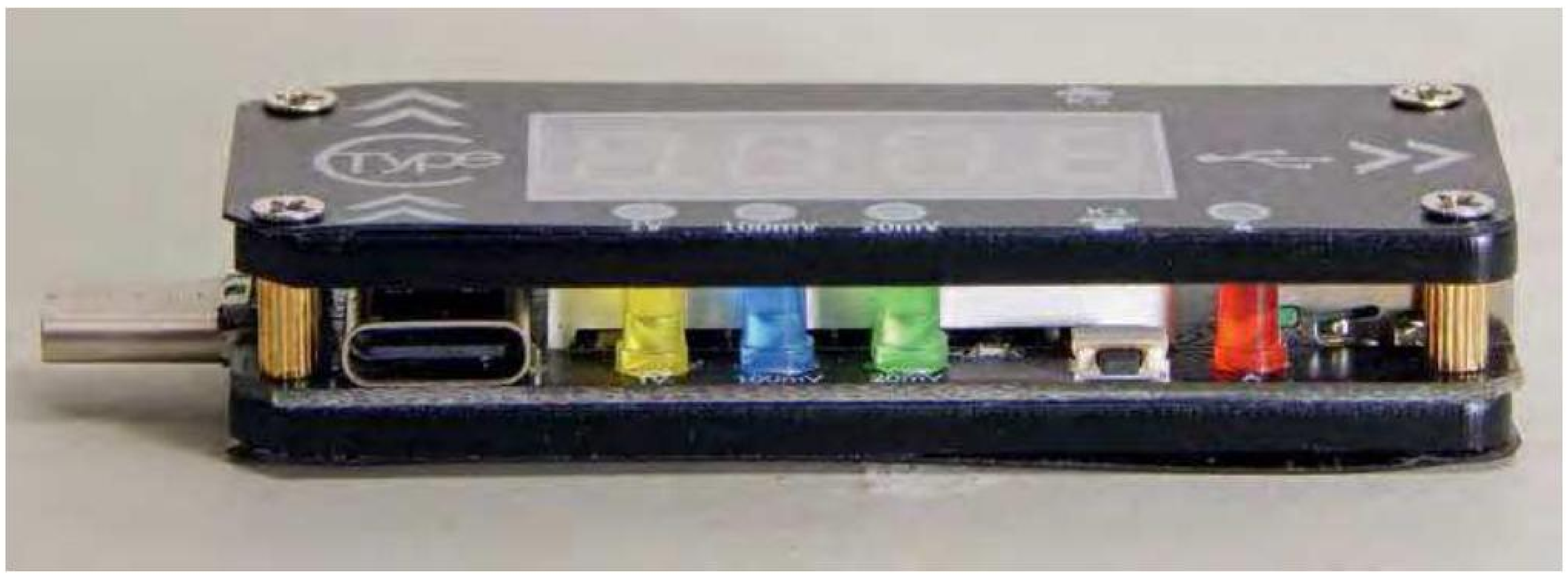
USB-C **https://w.wiki/nto**
USB-PD **https://w.wiki/34dT**
https://bit.ly/pe-sep22-usbc1
Quick Charge **https://w.wiki/34dU**



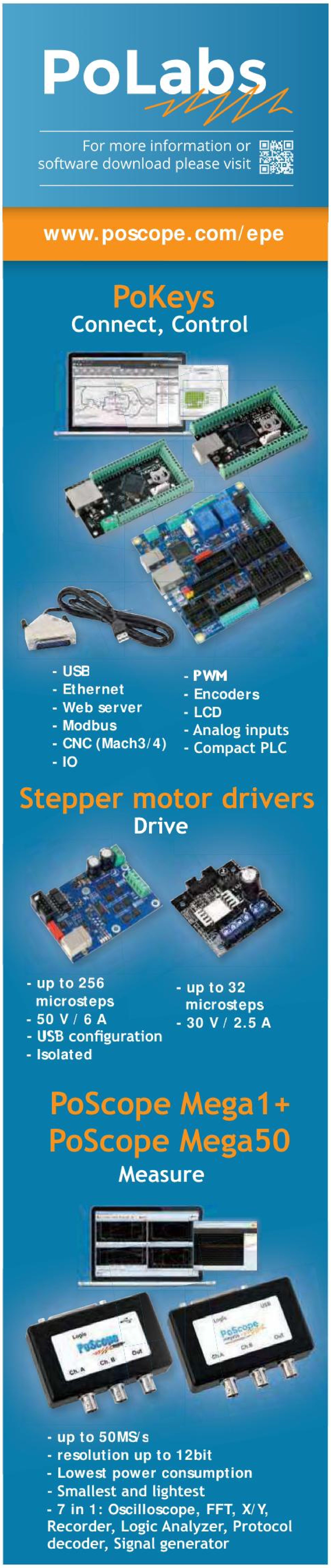
The XY-PDS100 quick charger (detailed last month) is shown connected to the XY-WPDT trigger unit, displaying the output voltage.



Here's what the assembled XY-WPDT module looks like. The two extra USB-C sockets on either side allow the unit to operate in pass-through mode.



From left-to-right we have the USB-C input, PPS mode LEDs (1V, 100mV and 20mV steps), K1 switch, and current indicator LED. The USB-C input is used with the matching output connector on the opposite side to control and measure voltage or current. Pressing the K1 switch changes between displaying current or voltage, while holding K1 just turns the screen and indicator LED off. Switch K2 is used in conjunction with K1 to change the voltage setting, and is a bit more complicated to set, see: https://bit.ly/pe-sep22-xy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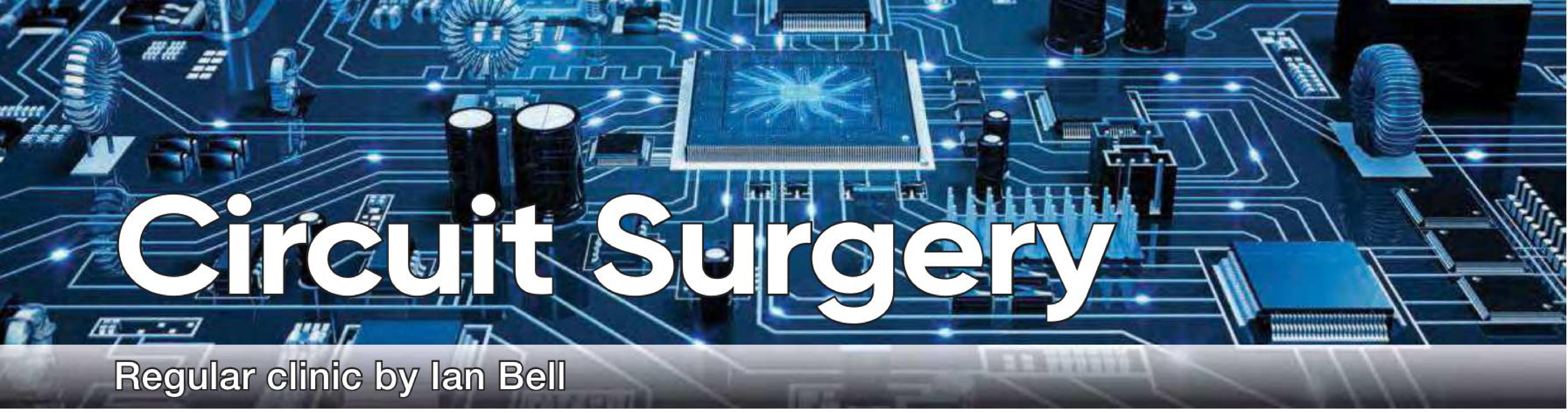


Circuit Surgery

Regular clinic by Ian Bell

Electronically controlled resistance – Part 1

Recently, on the EEWeb forum, user *Kintaro* posted a question about controlling resistance electronically. 'Right now, I have two values known, these are the voltage of a node (X volt) and a resistor (Y ohm). This circuit needs to interpret X volt as X ohm, and then tune the adjustable resistor to Z ohm (because the Y ohm resistance and Z ohm resistance in parallel give us the X ohm resistance. The resulting circuit should work for all the values of known values of the voltage node and the resistor.'

The key thing here is the ability to control a resistance with a voltage – the requirement for a parallel resistor is specific to Kintaro's application and can be achieved if the adjustable resistor can be set to the required value. We will look briefly at this and then consider electronically controlled resistance more generally.

Electronic control of resistance

There are a couple of widely used ways of achieving an electronically controlled resistor (other approaches are possible). First, field-effect transistors (FETs) can be used as voltage-controlled resistors. Both JFETs (junction field-effect transistors) and MOSFETs (metal-oxide-semiconductor field-effect transistors) can be employed for this purpose, but JFETs are quite common for this application. Second, there are numerous digital potentiometer ICs available. These typically contain a network of resistors and switches and are designed for use with microcontrollers, often via a serial bus such as SPI or I^{2}C. The digital instruction determines the state of the switches, which in turn control the resistance.

As with mechanical variable resistors, electronically controlled resistors can be either single resistors (sometimes referred to as a 'rheostats') or potentiometers (see Fig.1). A potentiometer has a 'wiper' terminal which connects to a variable location on a resistance or chain of resistors. The resistance of the potentiometer (between A and B in Fig.1) is fixed and the control (mechanical or electronic) sets the value of the resistance from A to the wiper, and from B to the wiper. This can be anything from zero to the A to B resistance, with the two summing to the A-to-B resistance. FETs naturally form single voltage-controlled resistors (rheostats), whereas digitally controlled resistors are available in both rheostat and potentiometer versions.

Simulation files

Most, but not every month, LTSpice is used to support descriptions and analysis in *Circuit Surgery*.
The examples and files are available for download from the *PE* website.

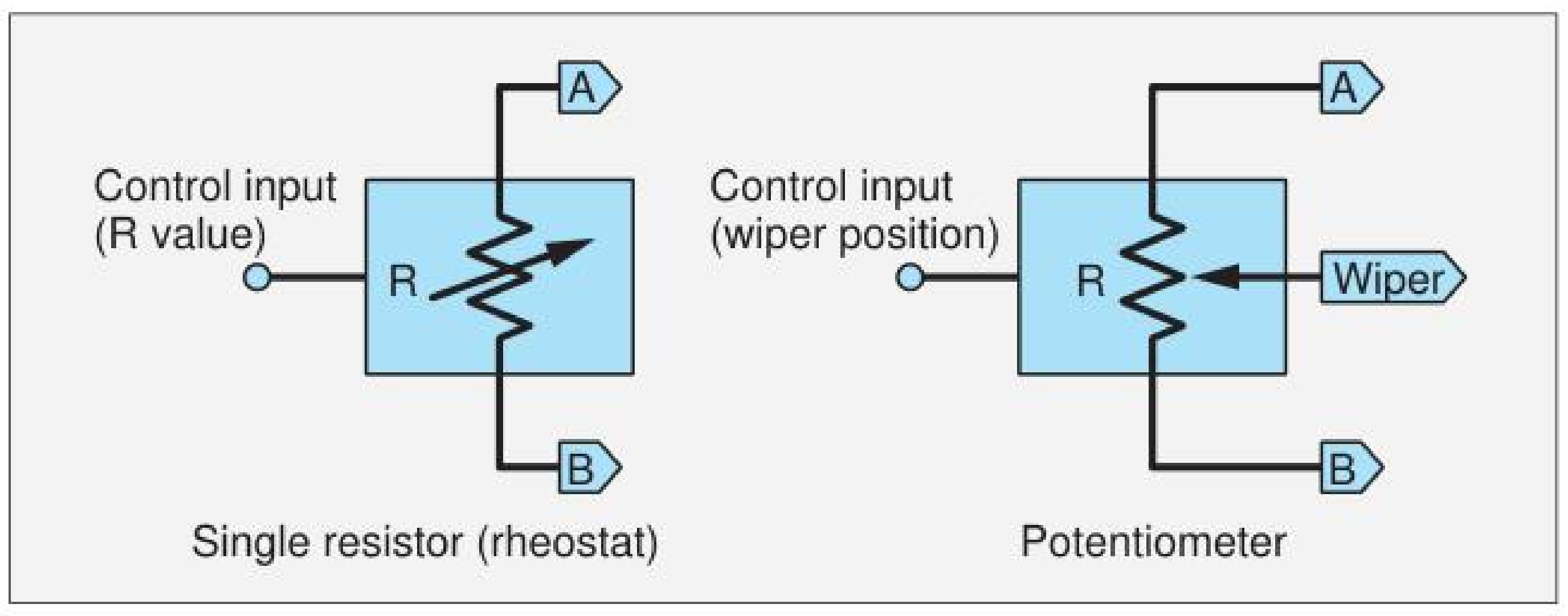


Fig.1. Electronically controlled resistors.

All components have limits to the range of voltages and currents that they can handle, but these may be more restricted for electronically controlled resistors than similar value mechanical devices. Electronically controlled resistors may have other restrictions such as the requirement for one end to be grounded.

Electronically controlled resistors could be considered in any application where a mechanical variable resistor, potentiometer or trimmer might be used but you want to facilitate digital (computer/microcontroller) control and thus increase a designs functionality. Voltage-controlled resistors are commonly used to provide analogue control of amplifiers' and filters' (gain and frequency) response in applications such as automatic gain control (AGC), adaptive filters and voltage-controlled oscillators.

Parallel resistors

Kintaro's EEWeb post describes a parallel combination of a fixed resistor with an electronically controlled resistor (see Fig.2). We will look briefly, and in very general terms, at this scenario. One of the best-known formulas in electronics is the total resistance R_T of a set of parallel resistors R_1, R_2, R_3 …R_N. The reciprocal of R_T is the sum of the reciprocals of parallel resistors:

$$\frac{1}{R_T} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \cdots + \frac{1}{R_N}$$

For just two parallel resistors (as in Kintaro's circuit shown in Fig.2) the total resistance is:

$$R_T = \frac{R_1 R_2}{R_1 + R_2}$$

In Kintaro's application one resistor is fixed (say R_1) and the other (R_2) is varied to obtain the desired R_T. We can rearrange the above equation to give the required R_2 value to achieve a specified total resistance R_T:

$$R_2 = \frac{R_1 R_T}{R_1 - R_T}$$

This is not a linear function – see Fig.3 – which shows the R_2 value required to obtain R_T in the range 1 to 9Ω with R_1 = 10Ω. If the original control voltage (X in Kintaro's post) is linear, and R_2 varies linearly with its control voltage, then it cannot be used directly to linearly control R_T (in Fig.2).

To achieve a linear control, it would be necessary to map the input control voltage (Kintaro's X) to the required control voltage for R_2. This could possibly be achieved by a microcontroller measuring voltage X (via an ADC) and finding the required resistor control voltage, either using a look up table, or via calculation with the above equation. A DAC could set the R_2 control voltage.

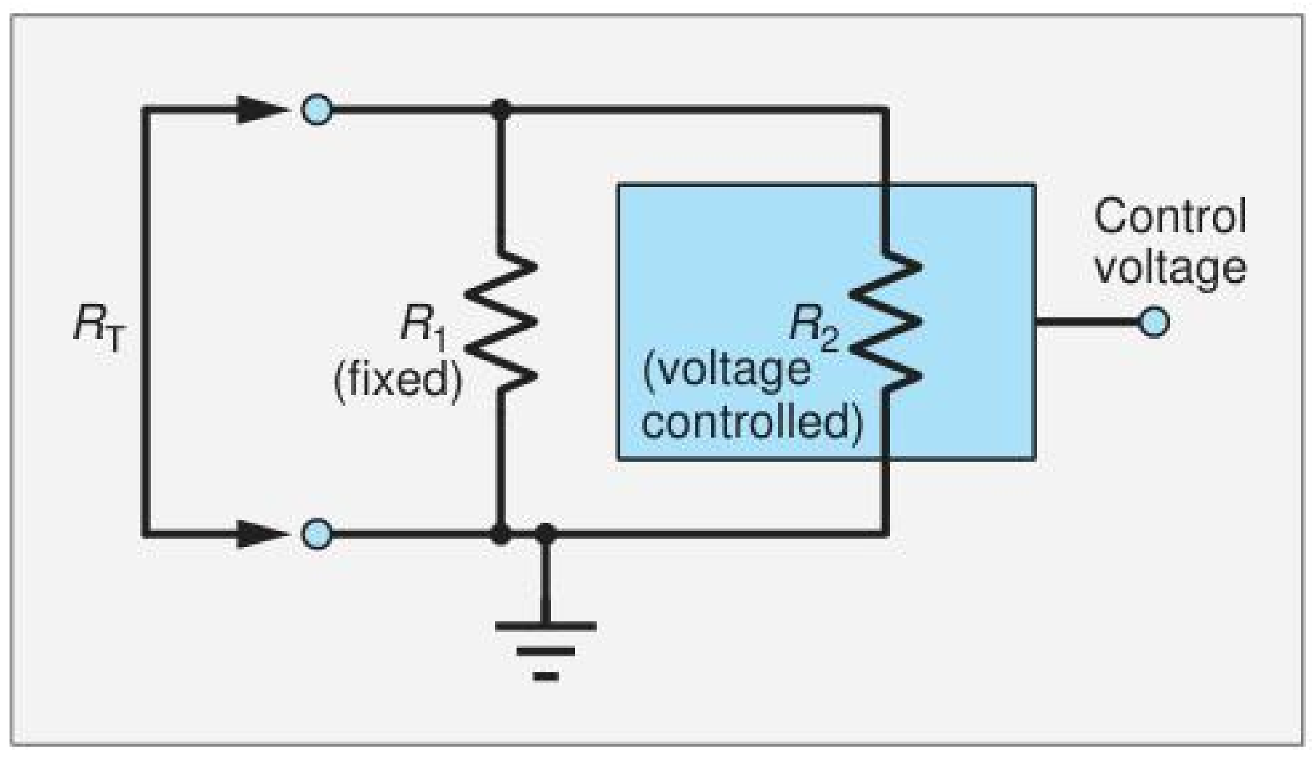


Fig.2. Fixed resistor R_1 in parallel with a voltage-controlled resistor R_2 giving a total resistance R_T.

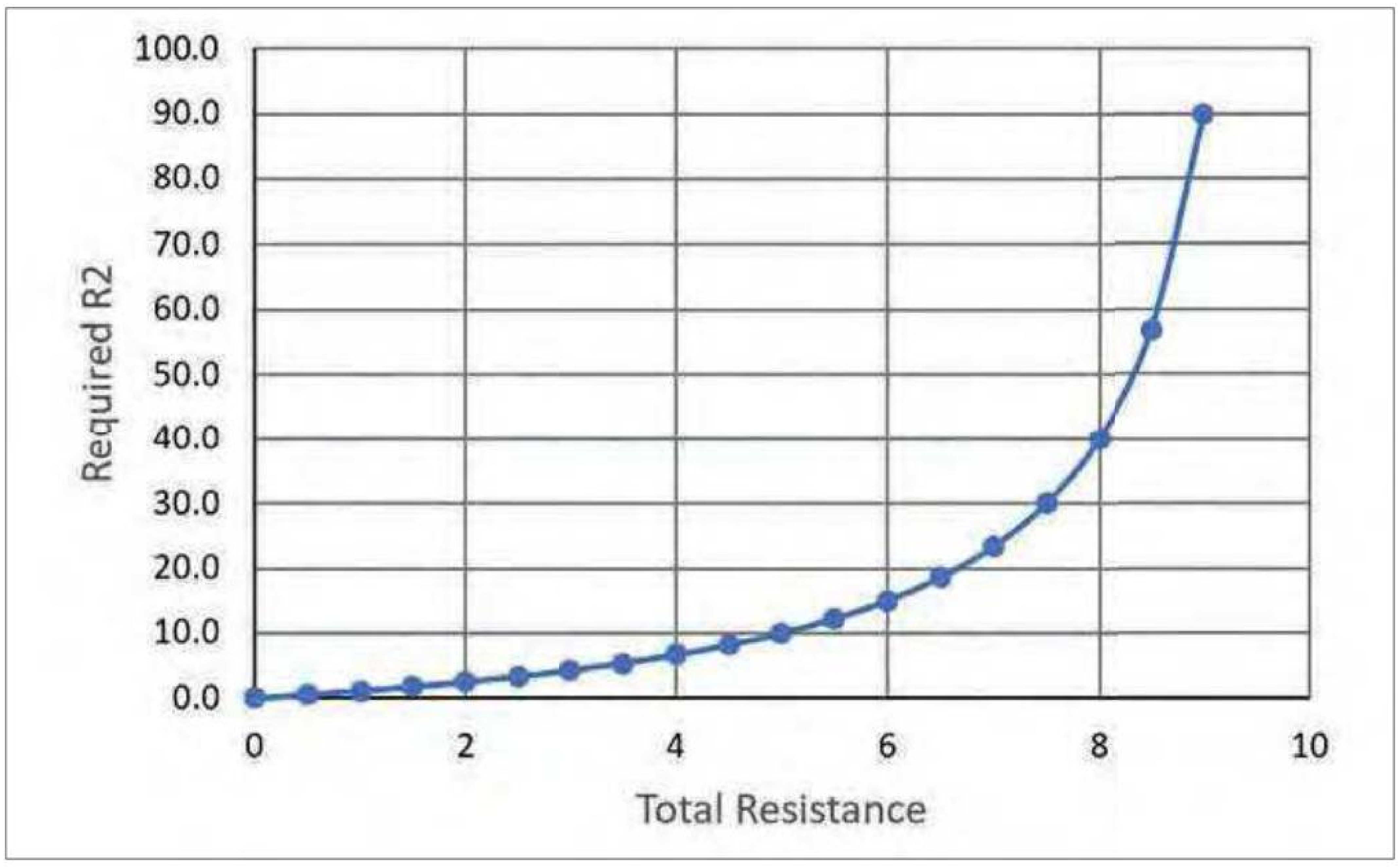


Fig.3. R_2 needed for the Fig.2 circuit to achieve the specified total resistance with R_1 =10Ω.

We do not know if the above solution is viable as we do not know the required response time, or any other details. Another issue is that the description of X volts to X ohms would imply low resistance values with typical circuit voltages in the units to tens of volts range and this might be difficult to achieve.

FETs as voltage-controlled resistors

As mentioned above, a JFET is commonly used to implement voltage-controlled resistance. The circuit in Fig.4 is an LTspice schematic which can be used to plot the characteristics of the device. JFETs have three terminals: gate, source and drain. The gate-source voltage (V_{GS}) (supplied by source `VGS` in the circuit in Fig.4) controls the drain current (I_D), which will also depends on the drain-source voltage (V_{DS}) (supplied by `VDS` in our schematic). The polarity of voltages and currents for a JFET depends on the type of device – n-channel or p-channel – a categorisation similar to the difference between NPN and PNP bipolar junction transistors.

The circuit in Fig.4 uses an n-channel JFET, which requires a negative gate-source voltage and typically a positive drain-source voltage, although, as we will see, *negative* drain-source voltages can also be used. The gate-source voltage controls the drain current, which flows into the drain. Note that the gate current is extremely small, so the JEFT is considered 'voltage controlled', unlike the bipolar transistor where we can consider either the base current or base-emitter voltage as controlling the collector current.

The LTspice simulation is configured to sweep both the drain-source voltage and gate-source voltage over a typical operating range, which results in the characteristics plot shown in Fig.5. This is similar to the characteristic plots you are likely to see in device data sheets.

The plot in Fig.5 is achieved using a DC sweep simulation (`.dc` SPICE directive) with two swept voltage sources. The first swept source is `VDS` – these voltages will be the *x*-axis of the plots produced by LTspice. The second swept source is `VGS` – each value this source takes will produce a separate curve on the plot. The plotted value is the drain current `Id(JI)`, which is selected for plotting in the same way as other types of simulation operation and is the *y*-axis of the graph.

The values `VDS 0 5 0.01` in the `.dc` directive (in Fig.4) specify that the `VDS` source (drain-source voltage) will be stepped from 0 to 5V in 0.01V steps. This produces a large number of datapoints on each curve which facilitates zooming in on the results. The values in the `.dc` directive `VGS 0 -2 0.25` specify that the `VGS` source (gate-source voltage) will be stepped from 0 to –2V in 0.25V steps – this is nine values, each of which will produce a separate curve on the plot (see Fig.5). (Note that we do not specify anywhere near as many steps on this second source sweep as with the first as too large a number would crowd the graph with an excess of curve plots.)

Fig.6 shows that we can divide the JFET's characteristics into two regions. The (right-hand) saturation region occurs at relatively high drain-source voltages and features a near constant drain current at a given gate-source voltage. This region is employed when the JFET is used as an amplifier. The other (left-hand) region is called the 'ohmic region' and is characterised by increasing drain-source voltages resulting in increasing drain current. This is more or less resistive behaviour, but, as is clearly seen in Fig.6 the current-voltage relationship is not linear (the plot lines are not straight throughout the ohmic region). However, if we look at relatively small drain-source voltages the lines are relatively straight – it is this part of the characteristic – within the arc

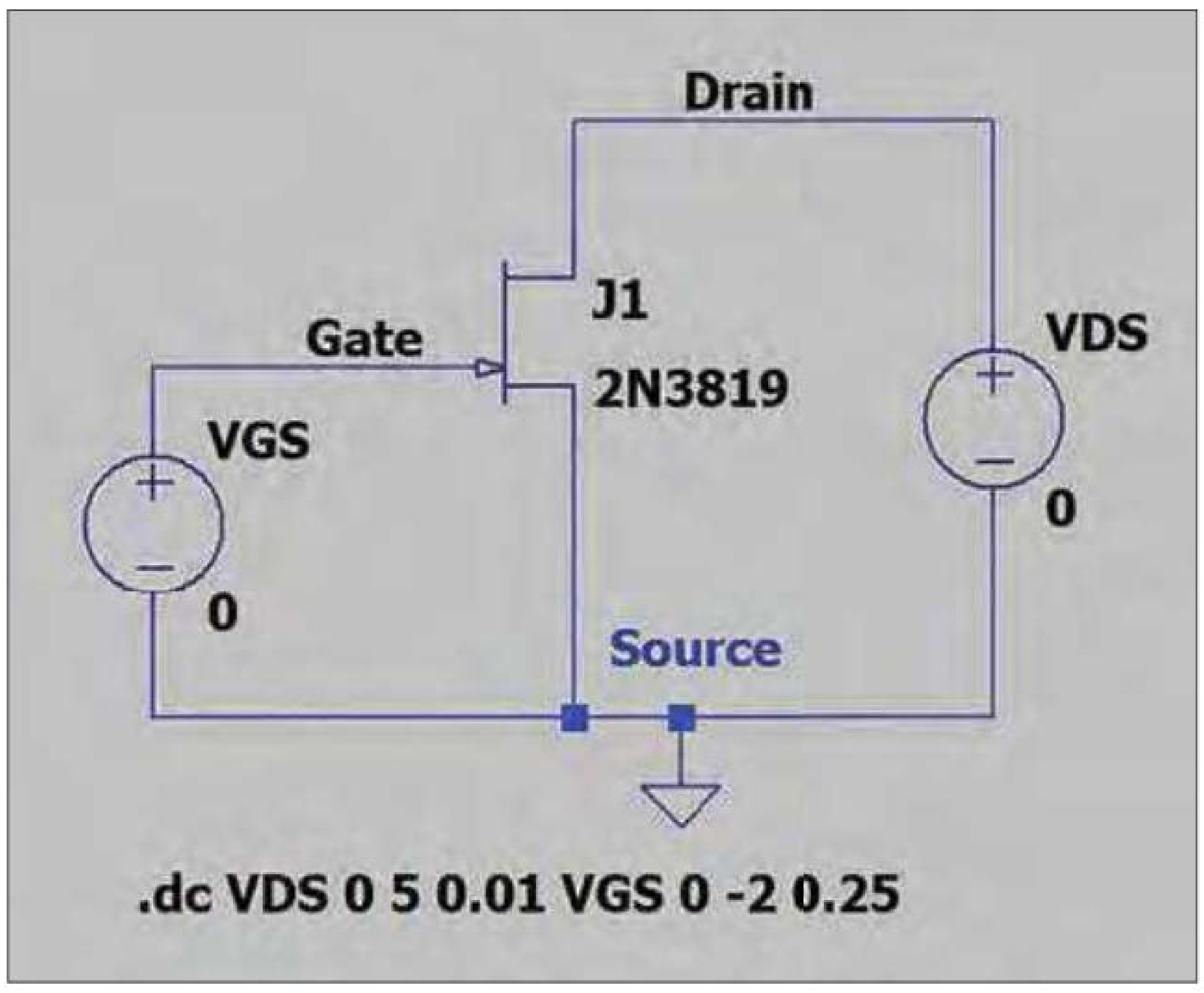


Fig.4. LTspice circuit for plotting JFET characteristics by varying V_{GS} and V_{DS}.

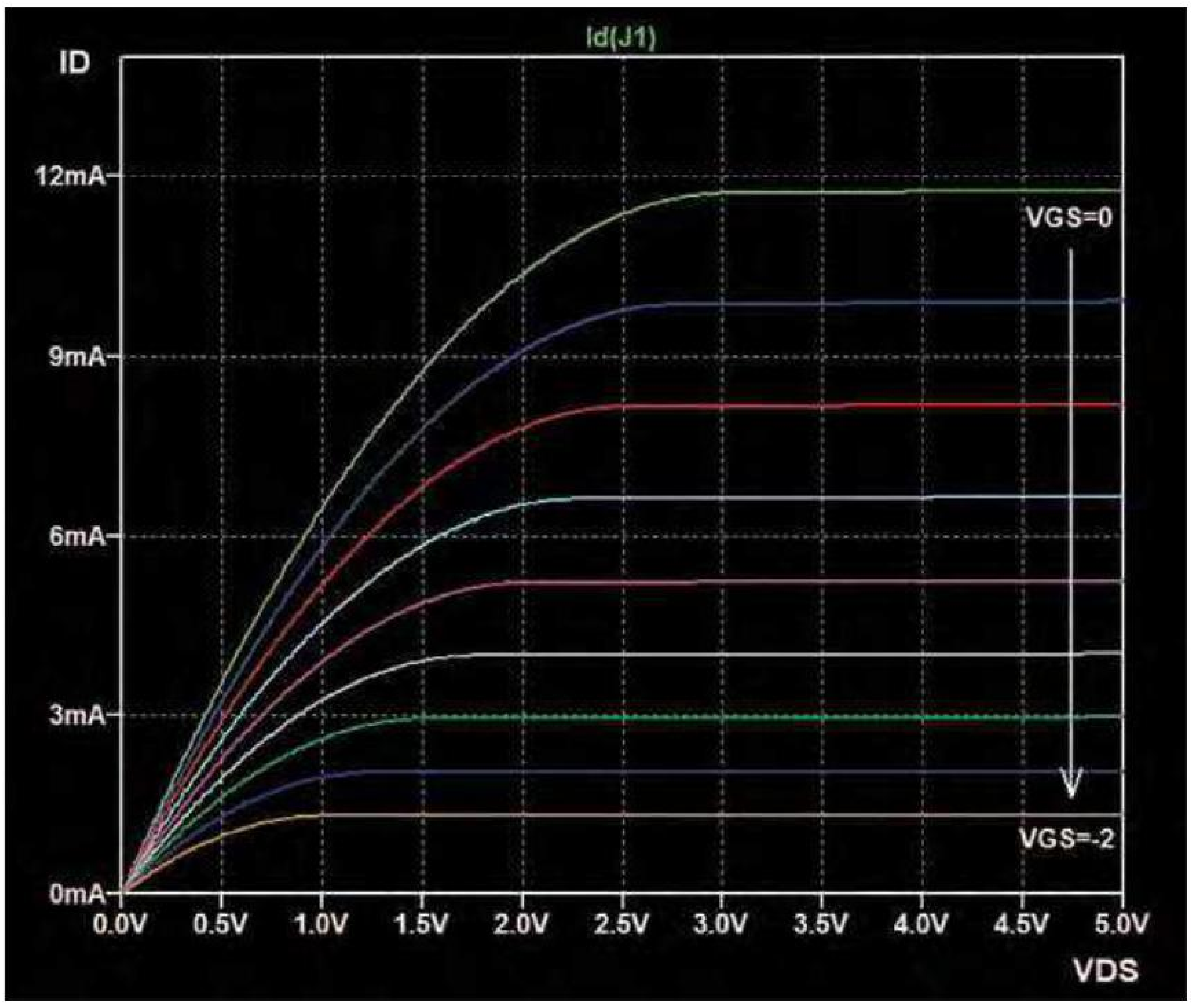


Fig.5. Typical JFET characteristics showing the 'ohmic' (left) and saturation (right) regions.

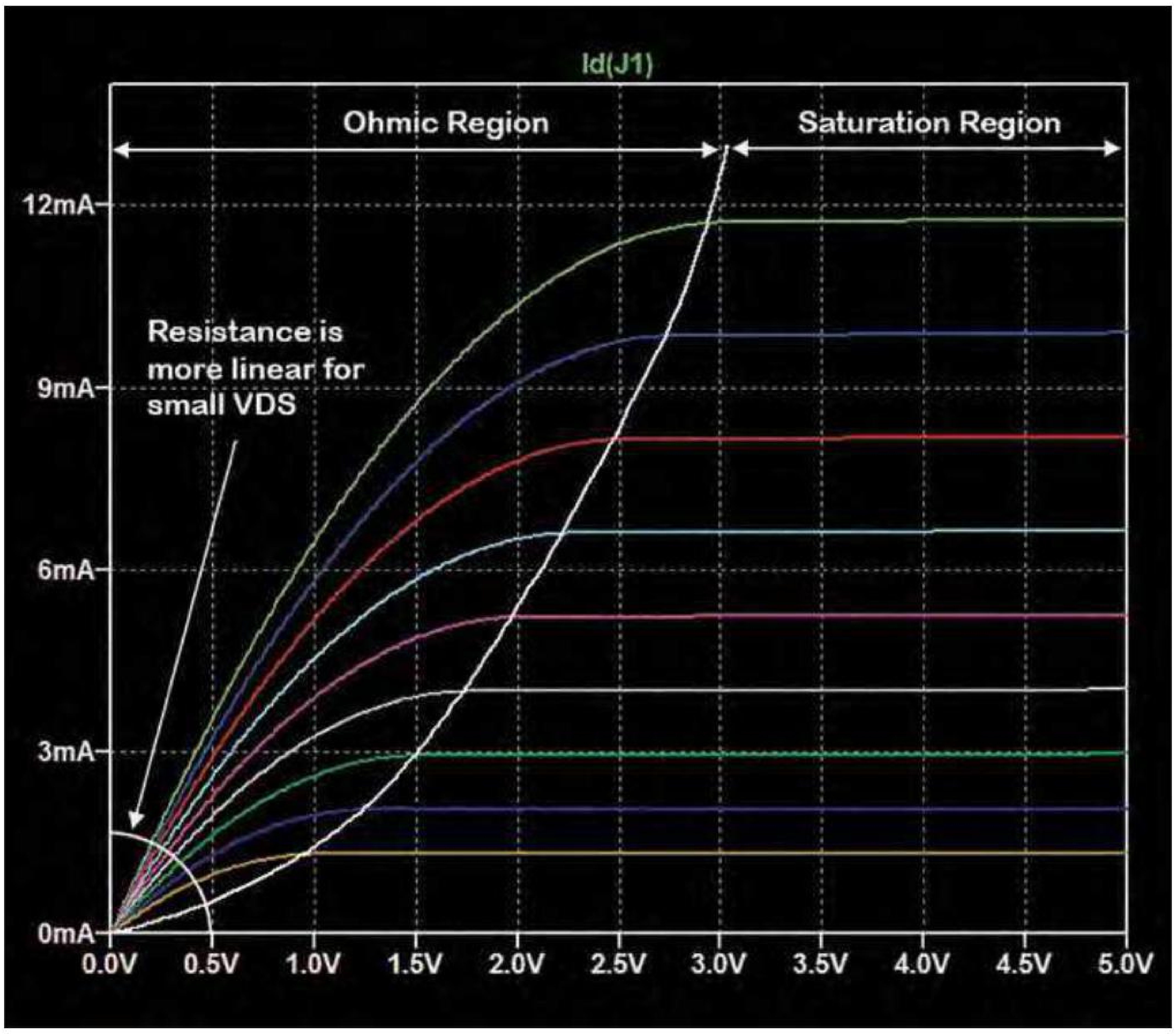


Fig.6. Regions of operation in JFET characteristics (I_D vs V_{DS} at various V_{GS}).

drawn near the origin that can be used to implement a voltage-controlled resistor.

Fig.7 shows a zoom in of the plot in Fig.6 to cover the region close to the origin. The graph has been extended to include negative drain-source voltages to illustrate the fact that a JFET voltage-controlled resistor works with voltages of either polarity across the 'resistor', that is, with either polarity of drain-source voltage. It can be seen that the current-voltage relationship in Fig.7 is fairly linear, but not perfectly so.

To get a better insight into the behaviour of the device as a voltage-controlled resistor it is useful to plot the resistance values directly. Ideally, these will be constant (flat line) and different for each gate-source voltage. LTspice is able to plot expressions based on circuit values, so we can plot drain-source voltage divided by the drain current to get the drain-source resistance. Specifically for the circuit in Fig.4 this is found using: `V(drain)/Id(J1)`. A plot of this is given in Fig.8. This shows the JFET's resistance varies from about 130Ω to 390Ω as `VGS` goes from 0V to –2V (at `VDS` = 0 V). The resistance is fairly constant over the plotted drain-source voltage range of ±600 mV and is better at lower magnitudes of gate-source voltage.

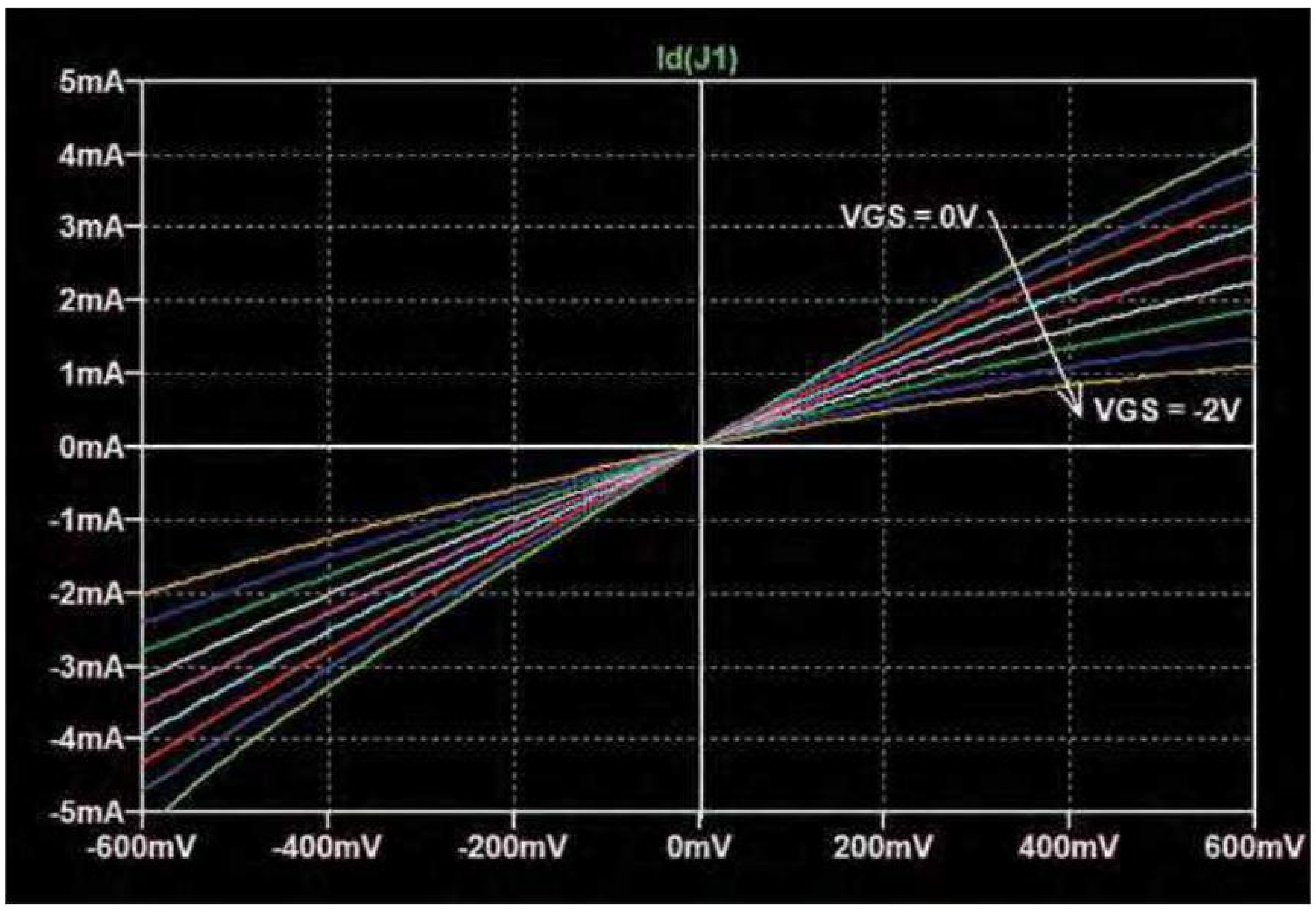


Fig.7. JFET characteristics for various gate-source voltages for drain-source voltages around zero.

The DC sweep command used to obtain Fig.7 and Fig.8 was slightly different from the one shown in Fig.4, specifically `.dc VDS -1.005 5 0.01 VGS 0 -2 0.25`. One change is that the `VDS` source sweep was started at near –1V so that we can plot voltages of both polarities. Also, the start of the sweep was at the odd-seeming value of –1.005V. This is to prevent a datapoint occurring at `V(drain) = 0`, because this results in zero drain current and hence the calculated resistance value is also zero. Although the actual resistance is not zero it cannot be calculated at a zero voltage, zero current point. This creates an anomaly in the plot that is avoided by the sweep values used. Note that although the plots in Fig.7 and Fig.8 are produced by zooming into LTspice, the large number of datapoints means that the curves are still accurate.

Fig.8 shows that the JFET's resistance is not completely constant with changing drain-source voltage. This will tend to cause distortion if the JFET 'resistor' is used in a signal path (we discussed distortion in detail in the last three *Circuit Surgery* articles). The JFET's resistance tends to increase as the drain-source voltage increases. It is possible to compensate for this by feeding back the drain-source voltage to increase the gate-source voltage at higher drain-source voltages. This will increase the drain current (with respect to a device with no feedback), resulting in a lower effective resistance.

A circuit to implement a more constant resistance is shown in Fig.9. The resistors used for the feedback must be large to prevent loading any circuit in which the JFET is used as a voltage-controlled resistor. 1MΩ is used in Fig.9, although we do not have any other circuitry to worry about. Hundreds of kilohms to megohms are typically used in such circuits.

Simulation results for the circuit in Fig.9 are shown in Fig.10 in a similar form to Fig.7 and Fig.8. The DC sweep was changed to account for the fact that the resistors `R1`

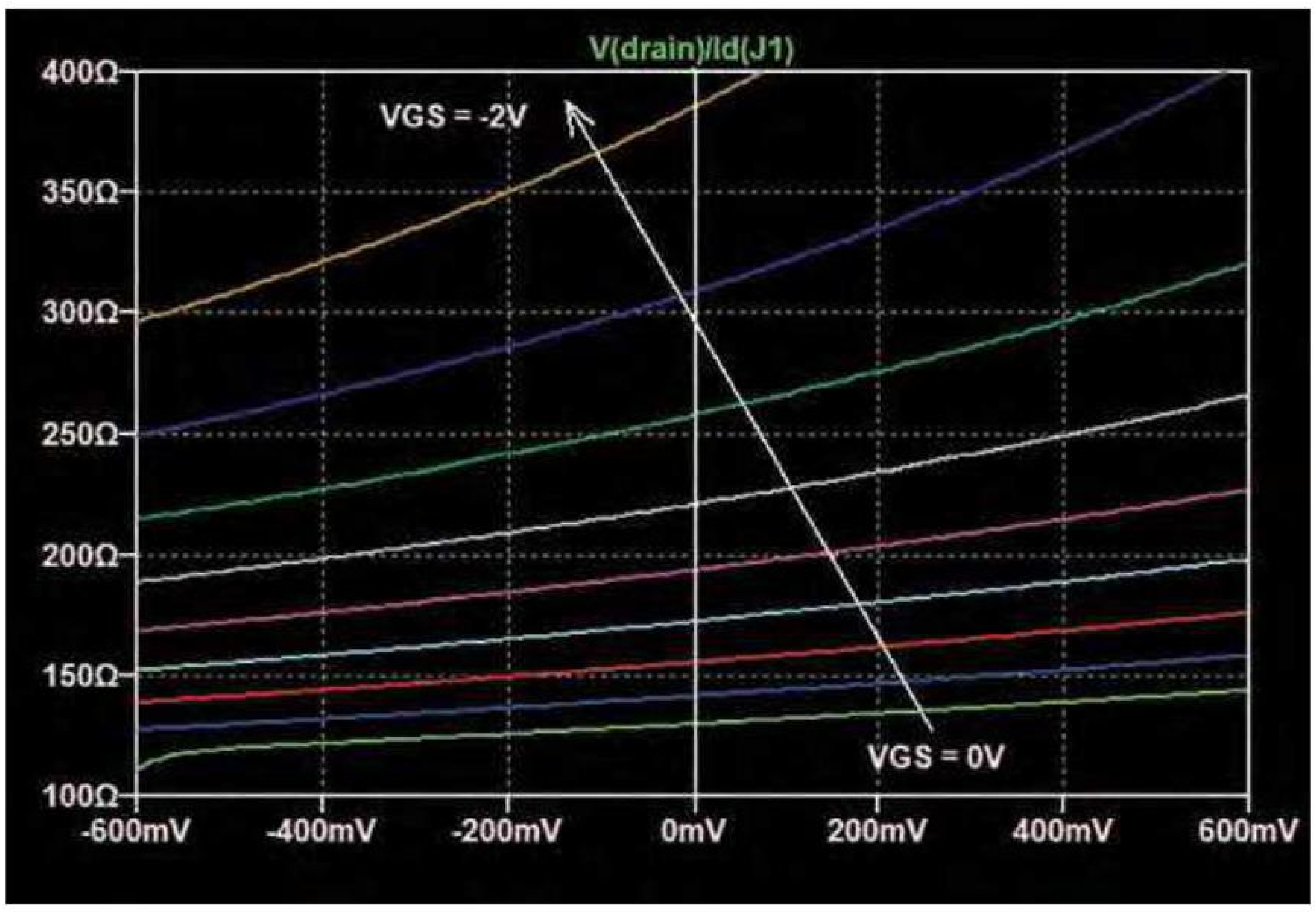
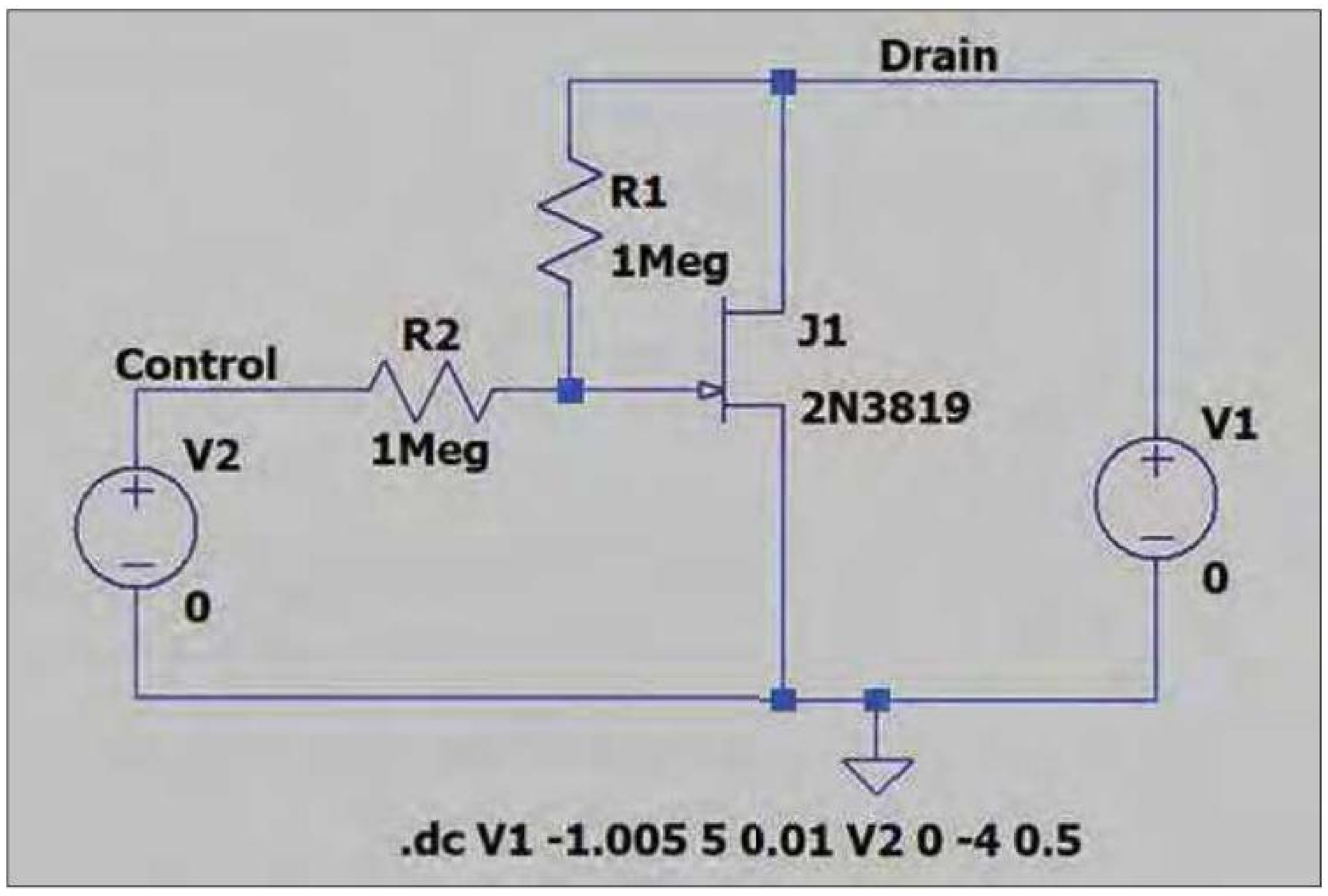


Fig.8. JFET drain-source resistance for various gate-source voltages for drain-source voltages around zero.



```
.dc V1 -1.005 5 0.01 V2 0 -4 0.5
```

Fig.9. JFET voltage-controlled resistor with feedback to give more constant resistance as drain-source voltage varies.

and R2 act as a divide-by-two voltage divider with respect to the effect of the control voltage (V2 source) on the gate-source voltage of the JFET. This voltage range and step size are doubled to –4V and 0.5V respectively to account for this. The results show a significant improvement in the linearity of the effective resistance – in other words, the resistance does not vary much with drain-source voltage (voltage across the 'resistor').

The JFET used in these examples was simply the first one in the list in LTspice. It was not chosen as an optimal device for voltage-controlled resistance applications but is sufficient to show the basic principles and the fact that any JFET can be used this way. JFETs optimised for use as voltage-controlled resistors are available, for example the VCR4N n-channel voltage-controlled resistor JFET from InterFET. These devices may well be (much) more expensive, for example, at the time of wiring a single 2N3819 is 93p from Mouser, whereas the VCR4N is £11.54.

Finally, Fig.11 shows one example of a circuit using a JFET voltage-controlled resistor. This is an amplifier with voltage-controlled gain. Resistor R1 and the voltage-controlled resistor formed by the JFET (J1) and the feedback resistors (R2 and R3) form a potential divider. The output level from this will depend on the controlled resistance value. It is then buffered and amplified by the op amp. The control voltage is negative (as in the circuits in Fig.4 and Fig.9), but this is not a major problem with a typical split supply for an op amp circuit. The input signal needs be centred on 0V and of sufficiently small amplitude to keep the JEFT circuit operating at a constant linear resistance at a given control voltage.

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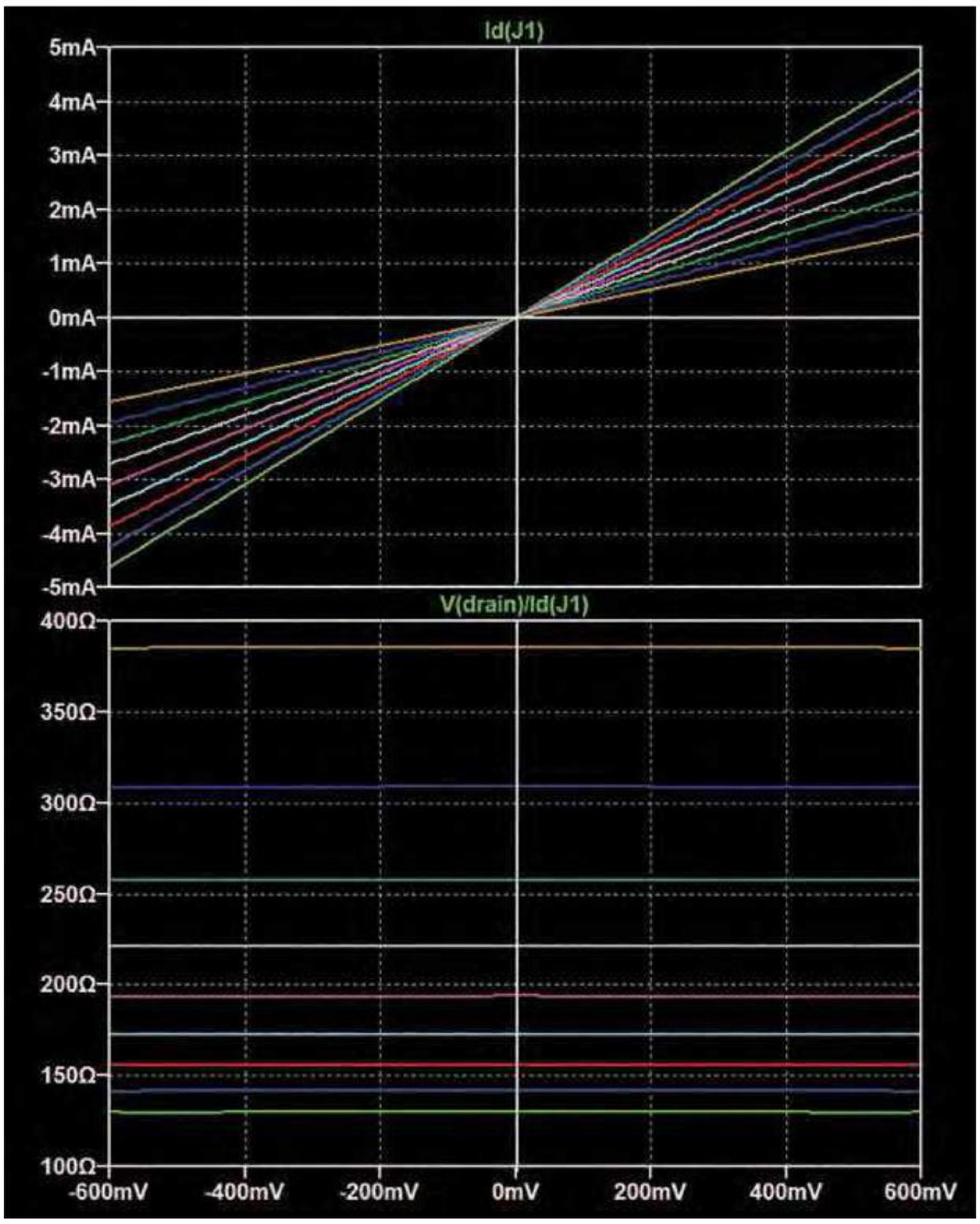


Fig.10. JFET characteristics and drain-source resistance with feedback applied as shown in the simulation circuit in Fig.9.

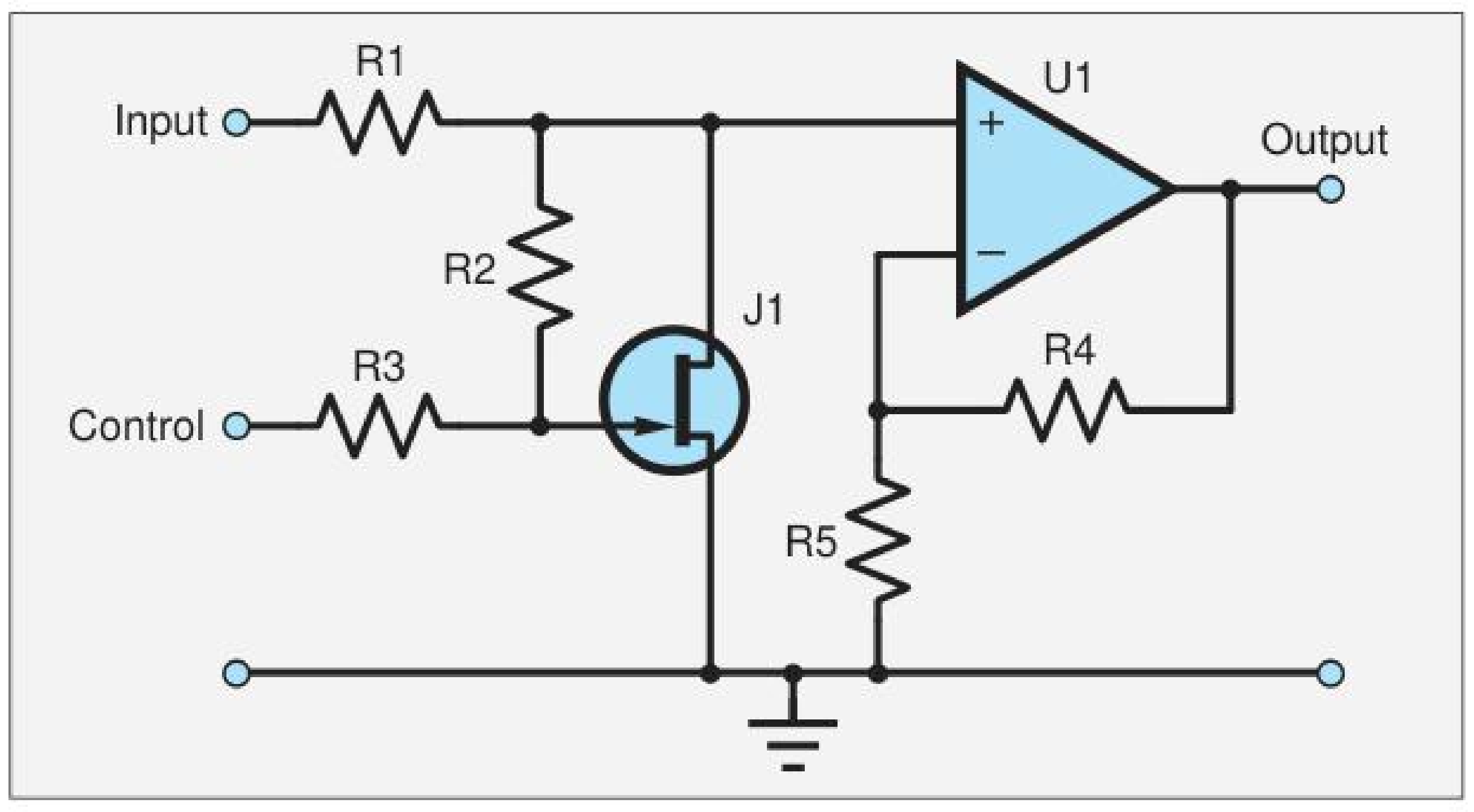


Fig.11. Amplifier with gain control using a JFET voltage-controlled resistor.

Make it with Micromite

Phil Boyce – hands on with the mighty PIC-powered, BASIC microcontroller

Part 39: Using infrared to synchronise time

Last month, we showed how easy it is to set up a *PicoMite BackPack*. We focussed on the 2.8-inch 320x240 touch TFT module, as this has the same-sized screen which was used in the original *Micromite BackPack* (and thus in many *PE* projects). In addition, we showed how to use the bigger 3.5-inch touch TFT with a higher resolution of 480x320. However, we also mentioned that there are many smaller (and cheaper) screens that can be used, which are handy if you only want to show a small amount of data, for example, the time. Since last month, we have received numerous emails about using these smaller screens; in particular, asking for help with the `OPTION` settings. Therefore, we will begin this month with a useful guide covering some of the more popular smaller screens.

While setting up the hardware for this task, I simply connected a PicoMite to a display, wrote a short program to show the time, and left the unit on test while setting up the next PicoMite. Fortunately, I had recently received a delivery of new RPi Pico modules and display modules; so I was able to set up multiple 'clocks' and position them side by side. Note that no RTC (real-time clock) was used in any of these clocks. Instead, each clock was configured using just MMBASIC's built-in `TIME$` variable. All I did was set `TIME$` at the command prompt to a value as close to the exact time as possible. Everything was going well, but then my 10-year-old son made a comment; he noticed that the clocks were all showing a slightly different time and asked, 'which clock is actually correct?'.

I gave the 'time is relative' answer, and then went on to tell him that probably none of the clocks were showing exactly the correct time. However, it gave me an idea for a topic that I had planned for a future article – infrared (IR) communication.

So, the second part of this month's article will demonstrate how to synchronise several clocks to show the same time by using MMBASIC's built-in `IR SEND` command. Essentially, we will show you how to add an IR transmitter to one clock (the 'master clock'), and then add an IR receiver to each of the other clocks so that they can all be set to show the same time.

For this article you will need at least two *PicoMite BackPacks* (ideally three or more), but do remember that a PicoMite costs less than £4, and displays start from only around £5 (and much less from the likes of eBay if you're prepared to wait for delivery from the Far East). The only other parts you will need are probably already in your spare parts draw: an NPN transistor, a couple of resistors, a single IR transmitter LED, and as many IR receivers as you have *PicoMite BackPacks*.

Micromite code

The code in this article is available for download from the September 2022 page of the *PE* website.

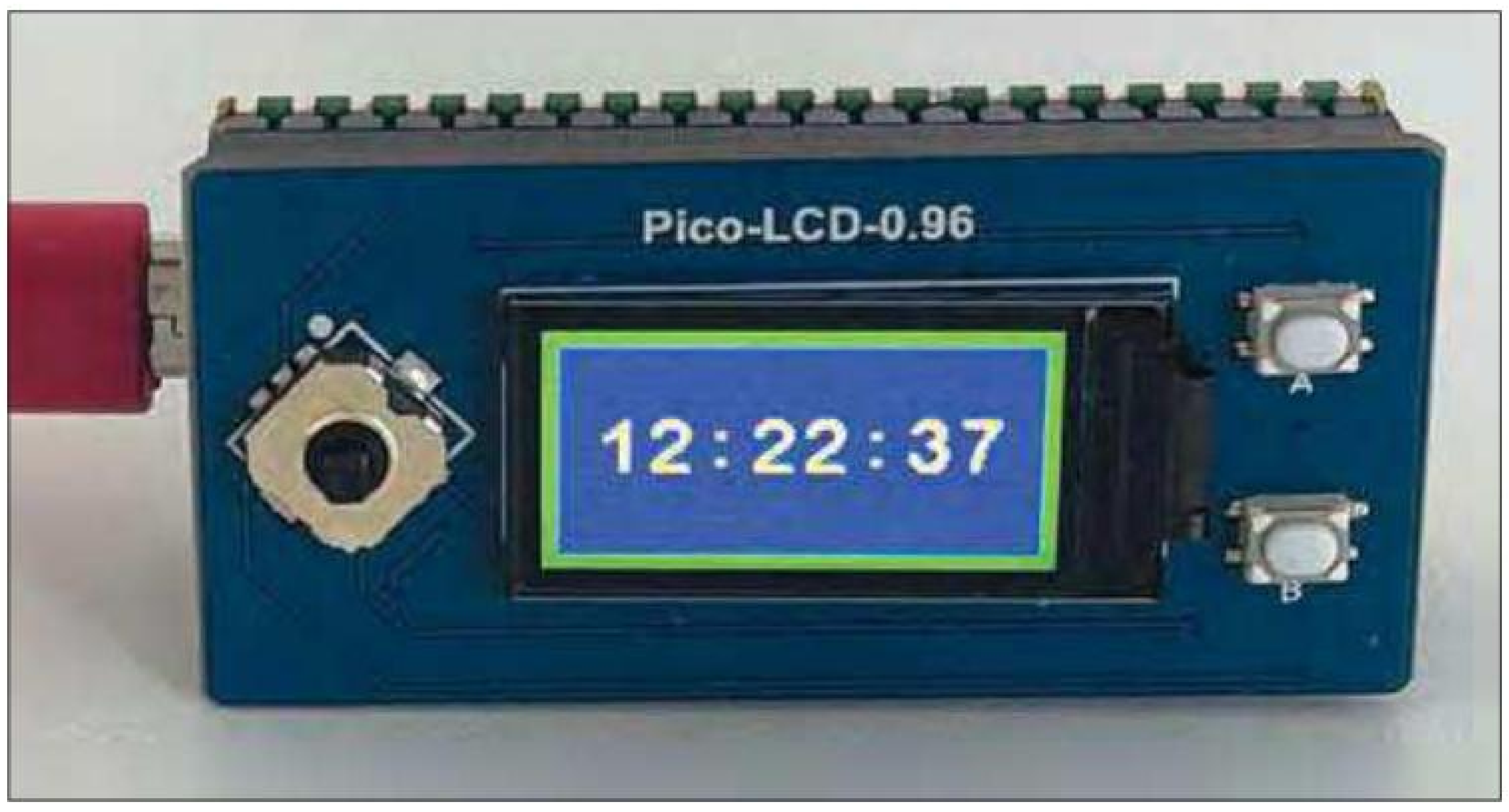


Fig.1. The Waveshare Pico-LCD-0.96 display module makes for a compact *PicoMite BackPack*. It has a really nice IPS screen with a resolution of 160x80.

Let's begin by showing the `OPTION` settings to create some *PicoMite BackPacks* with smaller screens.

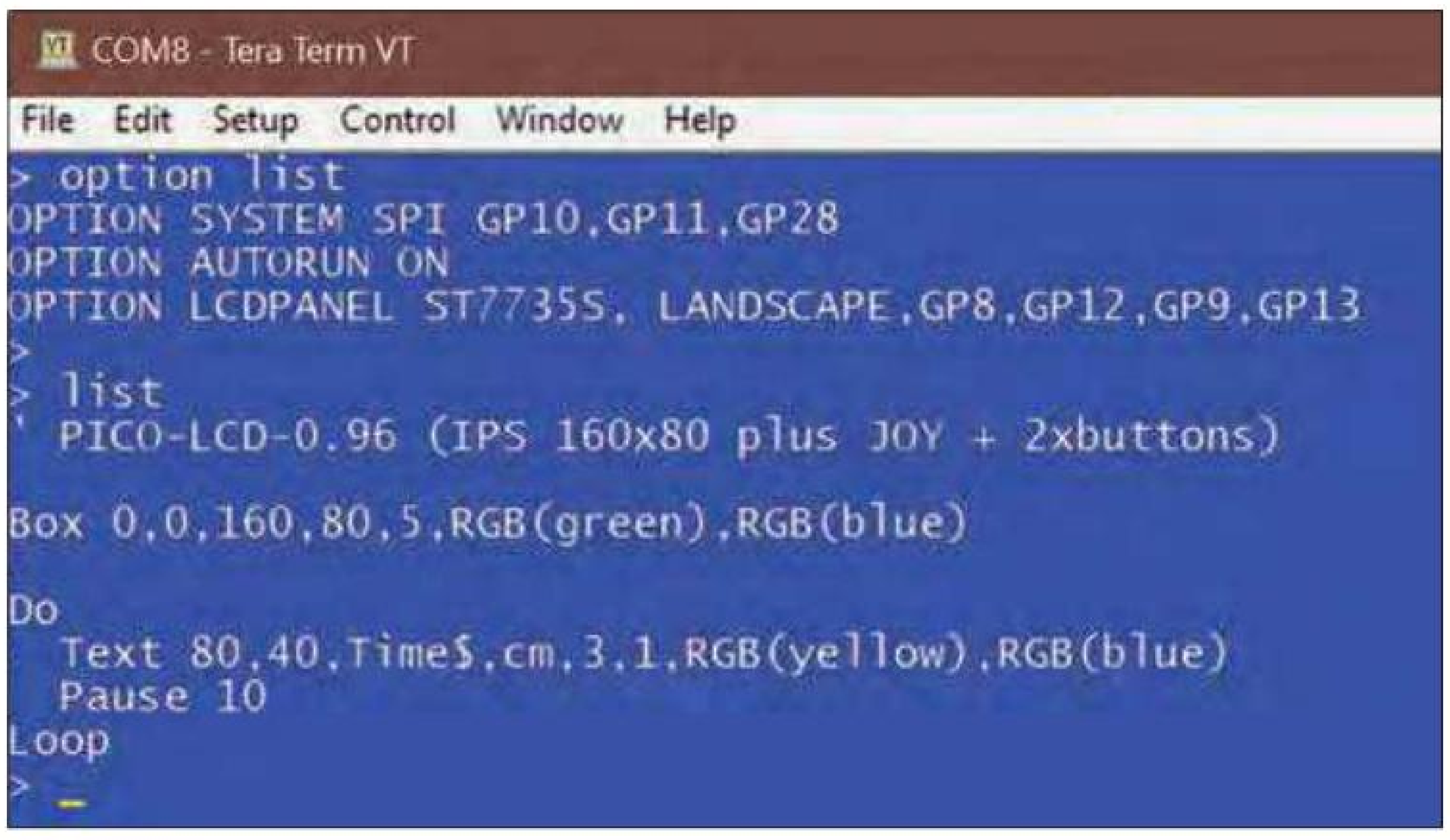
0.96-inch (160x80) IPS screen

This is one of my favourite small screens. It produces a very clear image with vivid colours, and a great viewing angle. It is available from many suppliers as a standalone screen (and is the same screen we used back in *MIWM*, *Part 9*, *PE*, Oct 2019). It is also available as a Pico plug-in module (see Fig.1 from last month – top-left image). If you search online for 'Waveshare Pico-LCD-0.96' you will see that this plug-in display module costs around £7.

With the display module and PicoMite connected to each other, you will need to enter these two configuration settings at the command prompt:

```
OPTION SYSTEM SPI GP10,GP11,GP28
OPTION LCDPANEL ST7735S,L,GP8,GP12,GP9,GP13
```

Once these are set, you will find all the usual MMBASIC graphical commands will work with this screen. To test this, try it with something like: `CLS RGB(CYAN)`. If you do not see the screen change to the appropriate colour, type `OPTION LIST` and check the parameter values are entered as shown above. Any errors will need to be fixed – you do this by first typing `OPTION LCDPANEL DISABLE` and then either `OPTION SYSTEM SPI DISABLE` (if you made an error with one of those



```
> option list
OPTION SYSTEM SPI GP10,GP11,GP28
OPTION AUTORUN ON
OPTION LCDPANEL ST7735S, LANDSCAPE,GP8,GP12,GP9,GP13
>
> list
' PICO-LCD-0.96 (IPS 160x80 plus JOY + 2xbuttons)

Box 0,0,160,80,5,RGB(green),RGB(blue)

Do
  Text 80,40,Time$,cm,3,1,RGB(yellow),RGB(blue)
  Pause 10
Loop
> _
```

Fig.2. Ensure your `OPTION`s are set with these parameter values to make the Pico-LCD-0.96-inch screen operate correctly. Shown here, the program listing for displaying the clock in Fig.1.

parameters), or by re-entering the `OPTION LCDPANEL` setting, this time with the correct parameters.

Next, set the `TIME$` variable as close as possible to the correct time with the command `TIME$="hh:mm:ss"` (replacing *hh*, *mm* and *ss* with the appropriate values). Finally, enter the following five-line program code:

```
BOX 0,0,160,80,5,RGB(green),RGB(blue)
DO
  TEXT 80,40,TIME$,cm,3,1,RGB(yellow),RGB(blue)
  PAUSE 10
LOOP
```

`RUN` the above program to see the time displayed in the centre of the screen; the result is shown in Fig.1. It is also worth making the program automatically start up on power-up, so stop the program (`Ctrl-C`), type `OPTION AUTORUN ON` at the command prompt, and then `RUN` the program again. Note that if power is removed, then the `TIME$` variable will be reset to '00:00:00' at power-up. This is useful because it will better demonstrate the IR time-synchronising that we will cover shortly. Fig.2 is a screen grab of the `OPTION` settings, and also shows the program listing (with a commented line at the start showing which screen it is).

1.3-inch (240x240) square TFT

Another useful Pico plug-in display module is the Waveshare PICO-LCD-1.3 (top-right, Fig.1 in last month's article). In addition to the 240x240 screen, there are four pushbuttons, and a digital joystick (up, down, left, right and select). The cost of this display module is around £8. To configure this display, enter the following configuration settings at the command prompt:

```
OPTION SYSTEM SPI GP10,GP11,GP28
OPTION LCDPANEL ST7789,L,GP8,GP12,GP9,GP13
```

You will notice that the settings are almost identical to the 0.96-inch screen above – the only difference being the display driver parameter: `ST7789` instead of `ST7735S`. This is something I like about the Waveshare display modules; they try to use consistent GP pin numbers for identical pin functions across their range.

With the 1.3-inch display configured properly, quickly test it: set `TIME$` and enter the following five-line program:

```
BOX 0,0,240,240,5,RGB(green),RGB(blue)
DO
  TEXT 120,120,TIME$,cm,5,1,RGB(yellow),RGB(blue)
  PAUSE 10
LOOP
```

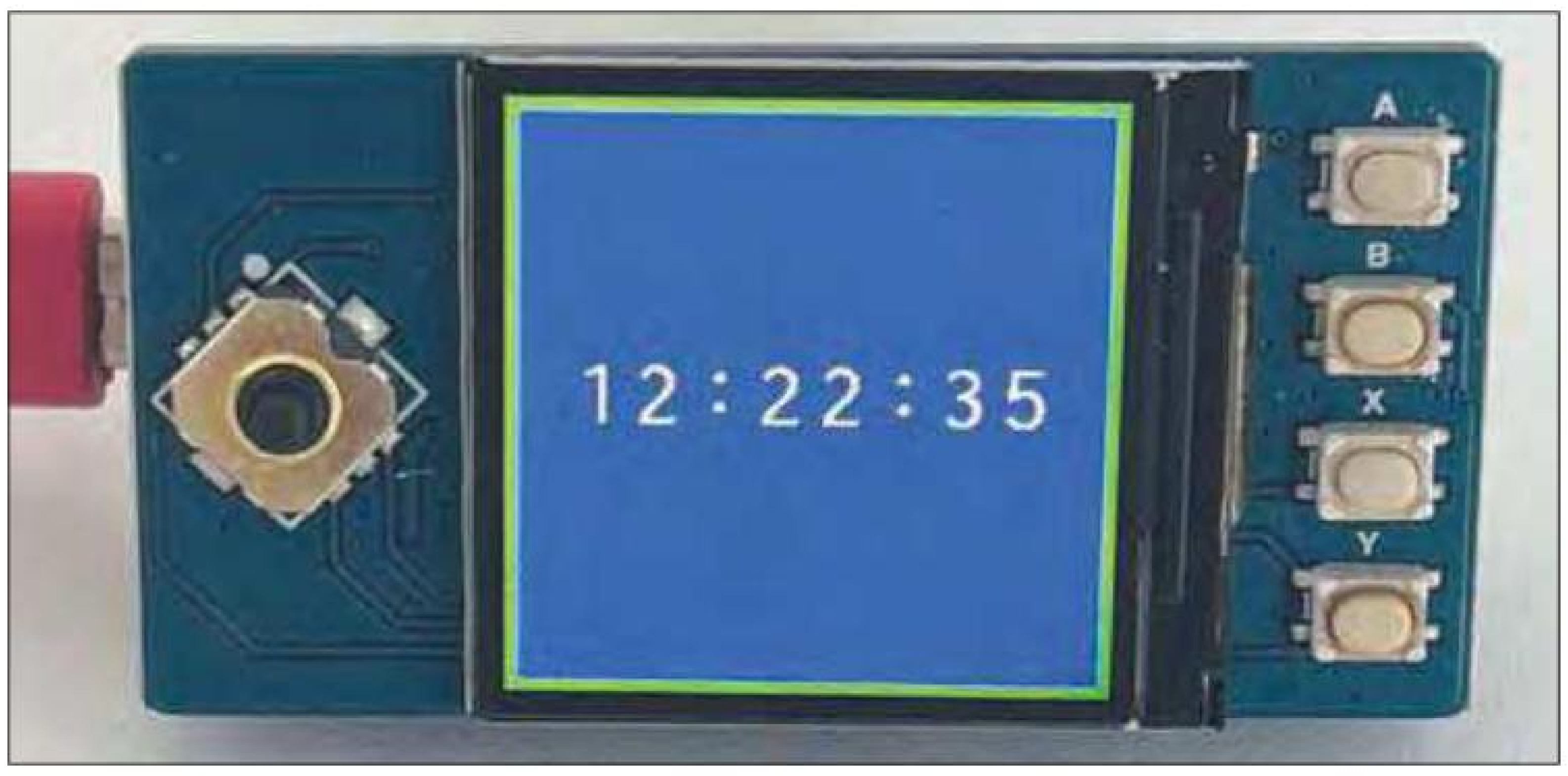


Fig.3. The Waveshare Pico-LCD-1.3 display module running the clock code, as listed in the text.

The program is identical to the one above except for the differences highlighted in bold. Because these two screens have different pixel resolutions the code needs to be altered so that items are displayed in the correct positions (and in a suitably sized font). With the program entered, set `OPTION AUTORUN ON` and then `RUN` the program. You will once again see the time displayed in the centre of the screen (see Fig.3), and that completes the second clock.

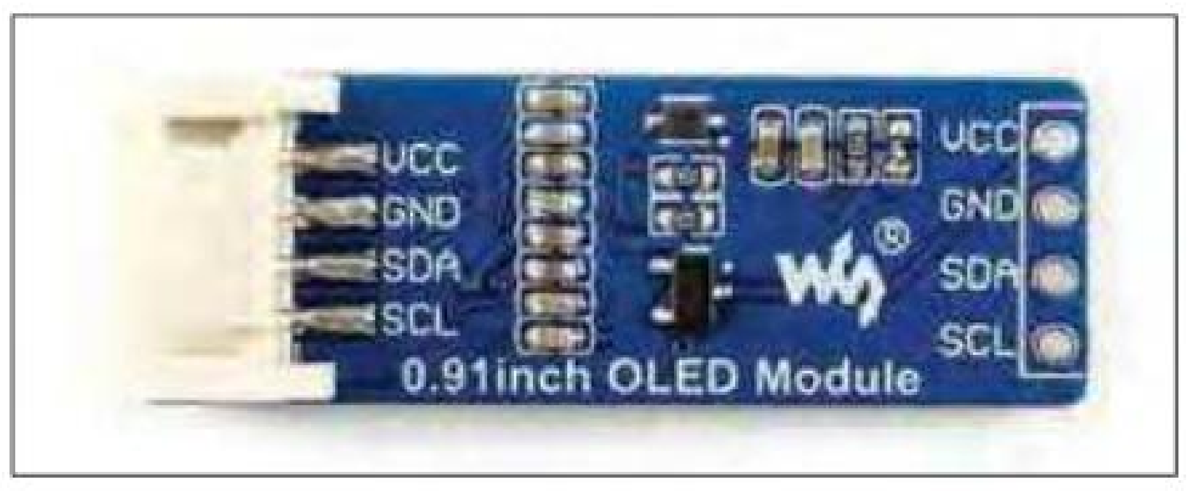


Fig.4. A standalone Waveshare 0.91-inch OLED module requires just four connections to the PicoMite – see text for details.

0.91-inch (128x32) OLED

OLED (organic light-emitting diode) displays are made up of individually controlled LEDs, so they don't need a backlight. Hence images on an OLED are generally superior to those on an LCD. They have high contrast, making them easy to read, even compared to the older-style LCD character modules, such as the 16x2, 20x2 and 40x4 displays.

Several Pico plug-in OLED display modules are available, but the current ones are not directly supported by the PicoMite because the relevant display drivers are not built into MMBASIC. However, there is nothing to stop us from using a standalone OLED module that is supported by MMBASIC. For example, if you search for 'Waveshare 0.91-inch OLED' you will see a compatible OLED module that costs around £5. Note that this OLED module is not a colour OLED; instead, it is mono (single colour), typically white or blue. This display is easy to interface to a PicoMite since there are just four wires to connect (refer to Fig.4):

OLED	PicoMite
VCC	3V3
GND	GND
SDA	GP0
SCL	GP1

To configure this OLED module, enter the following configuration settings at the command prompt:

```
OPTION SYSTEM I2C GP0,GP1
OPTION LCDPANEL SSD1306I2C32,L
```

As before, set the `TIME$`, set `OPTION AUTORUN ON` and enter the following program code:

```
CLS
DO
  TEXT 64,16,TIME$,cm,3,1
  PAUSE 10
LOOP
```

Once again, the time is displayed in the centre of the screen; or more realistically, due to the lower pixel-count (and appropriate font size), the time essentially fills the entire screen space – see Fig.5.

1.28-inch (240x240) round LCD

This screen provides a nice alternative to the usual square/rectangular displays. It is available as a Pico plug-in module – see Fig.1 from last month – top centre image. That one is advertised as an LCD screen; whereas the standalone round display is advertised as being an IPS screen (IPS is generally regarded as giving a much better display than an LCD). I ordered a standalone IPS display, but having seen it in operation, I must question whether it is really is an IPS display because the viewing angles are more like a standard LCD. Anyway, since I received several emails asking about the specific configuration setup for this display we'll continue. Bearing in mind I have a standalone display rather than a Pico plug-in module, I had to first make the following connections (refer to Fig.6):

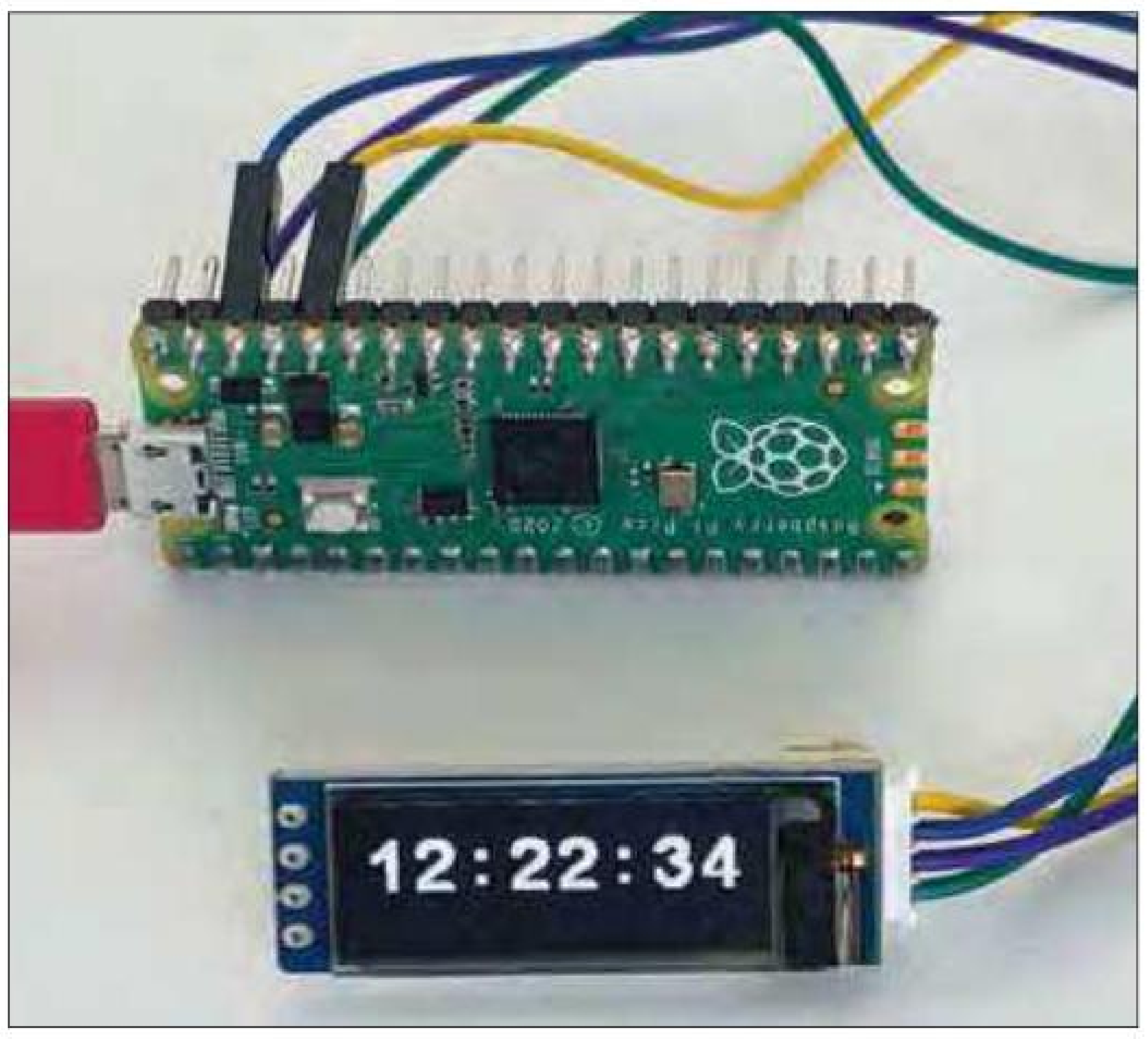


Fig.5. The Waveshare 0.91-inch OLED PicoMite clock – here it is using a white display.

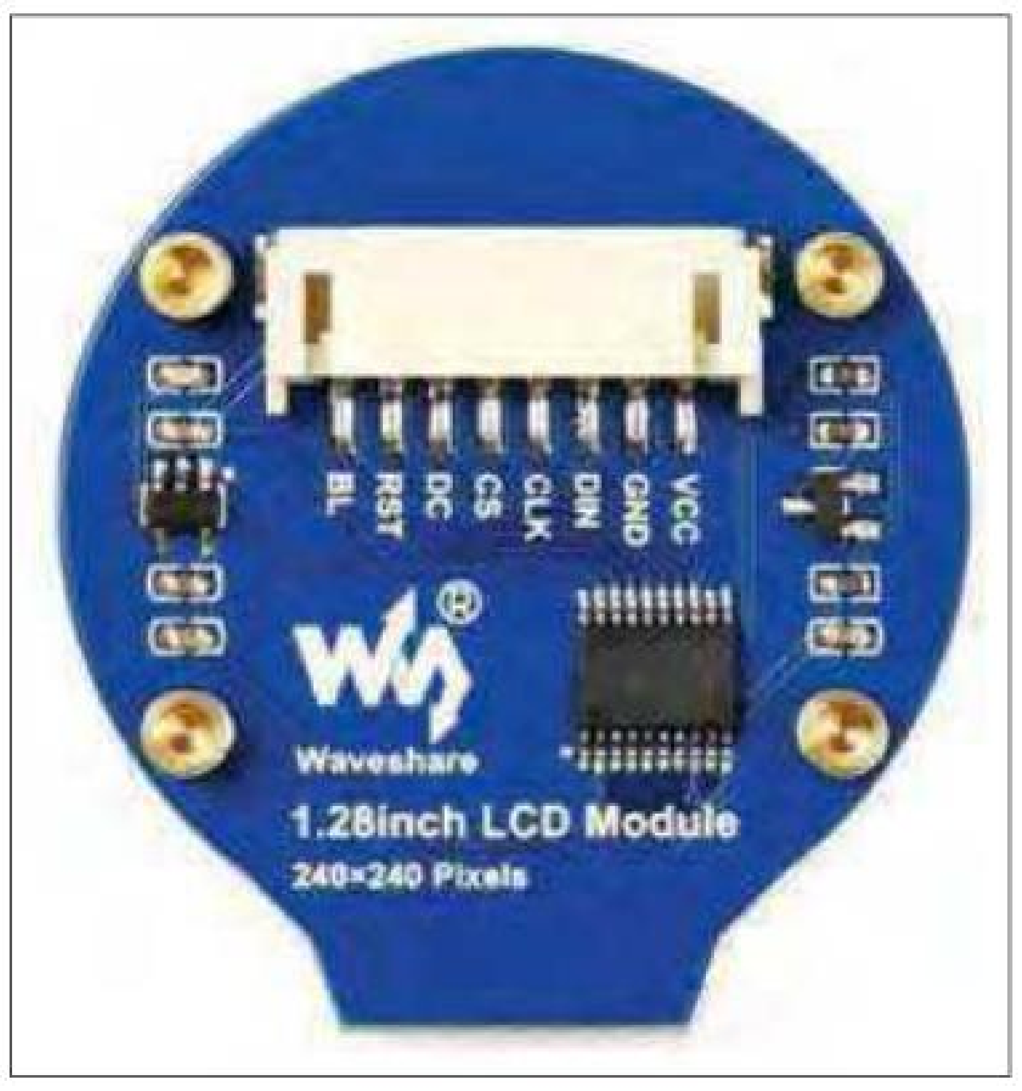


Fig.6. The Waveshare 1.28-inch round LCD standalone module needs eight connections to the PicoMite – see text for details.

LCD	PicoMite
VCC	3V3
GND	GND
DIN	GP11
CLK	GP10
CS	GP9
DC	GP8
RST	GP12
BL	GP13

The above pins were selected to match the pin-out of the equivalent Pico plug-in display module, so the following should work for either type of round display module. If ypu use a Pico plug-in module, you will notice that once again Waveshare have used consistent GP pin numbers for identical pin functions and hence it is just the display driver reference (GC9A01) that is the main difference from what was used on the previous (non-OLED) screens above.

By the way, the typical cost of the standalone display module is around £13, whereas the Pico plug-in module (complete with digital joystick) costs around £23.

To configure this display, enter the following configuration settings at the command prompt:

```
OPTION SYSTEM SPI GP10,GP11,GP28
OPTION LCDPANEL GC9A01,L,GP8,GP12,GP9,GP13
```

After configuring it, quickly test it with something like `CLS RGB(red)`, then set the `TIME$`, and also set `OPTION AUTORUN ON`. Last, enter the following program code:

```
CLS RGB(blue)
DO
  TEXT 120,120,TIME$,cm,2,2,RGB(yellow),RGB(blue)
  PAUSE 10
LOOP
```

The program is identical to that above except for the differences highlighted in bold. With everything entered correctly, you will once again see the time displayed in the centre of the round screen (see Fig.7); and that completes the fourth clock.

Adding IR

We are now going to discuss how to add IR functionality to the clocks so that we can demonstrate how to synchronise the clocks to show the same time. Depending on how many PicoMite clocks you have access to, you will need to modify the hardware of one (and only one) of them to act as the master clock by adding a simple IR transmitter (Tx) circuit. The remaining clocks are modified to act as slave clocks by adding an IR receiver (Rx) to each one.

You will also need to make suitable changes to the software. The master clock program code needs to be able to transmit it's time data via the IR Tx; and the slave clock program code needs to be able to receive this time information (via IR) so that it can update it's `TIME$` variable to the value received. Using this simple concept means that all clocks will display the same time.

Master clock

The master clock IR transmitter circuit is shown in Fig.8 – it can be connected to any spare GP output pin – in this case it is connected to GP22. The component values used do not need to be exactly the same – the circuit is provided more for guidance, but remember that a Pico output pin has very limited current drive capability, so an IR Tx LED cannot simply be connected to a GP output pin. With this circuit in place, the Pico remains within operational specification limits, and it also means that the Pico can drive the IR Tx at 5V (resulting in better range – not that much range is really required for this application).

Now build the IR Tx circuit and connect it to one of your PicoMite clocks. As a reference, we built the circuit on a breadboard, then used three male-female DuPont leads to connect it to the Pico-LCD-0.96 PicoMite Clock (that was moved onto an Expander module – see Fig.9). By using this display module for the master clock, we were able to use the bottom right-hand button (referred to as 'User Key B') as a trigger to transmit the time data. All that was required to achieve this was the command `SETPIN GP17,INTL,btn,PULLUP` at the start of the program code, meaning that the program will jump to a subroutine (called `btn`) whenever the button is pressed. We then put the IR transmitting

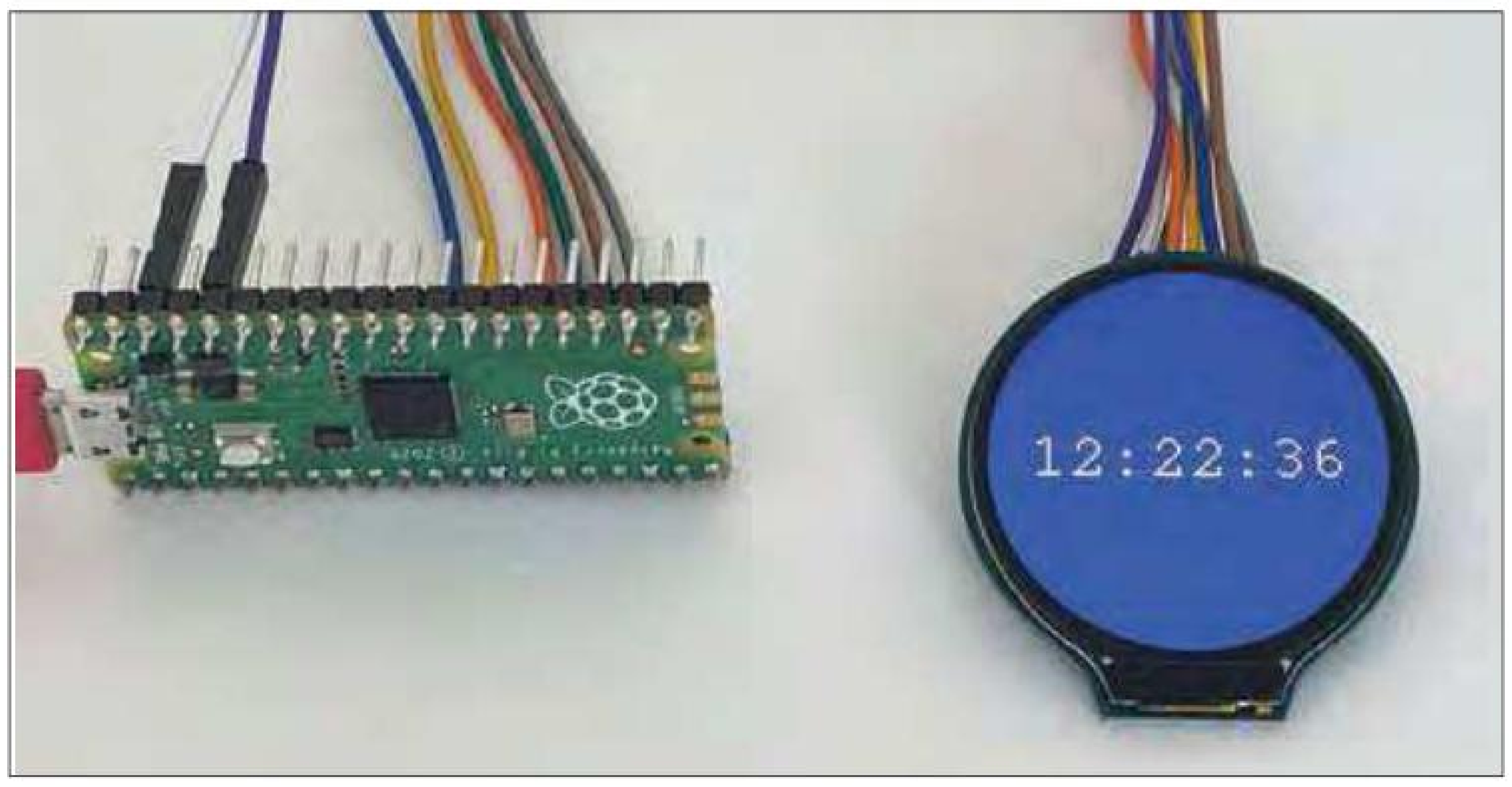


Fig.7. The Waveshare 1.28-inch round LCD PicoMite clock.

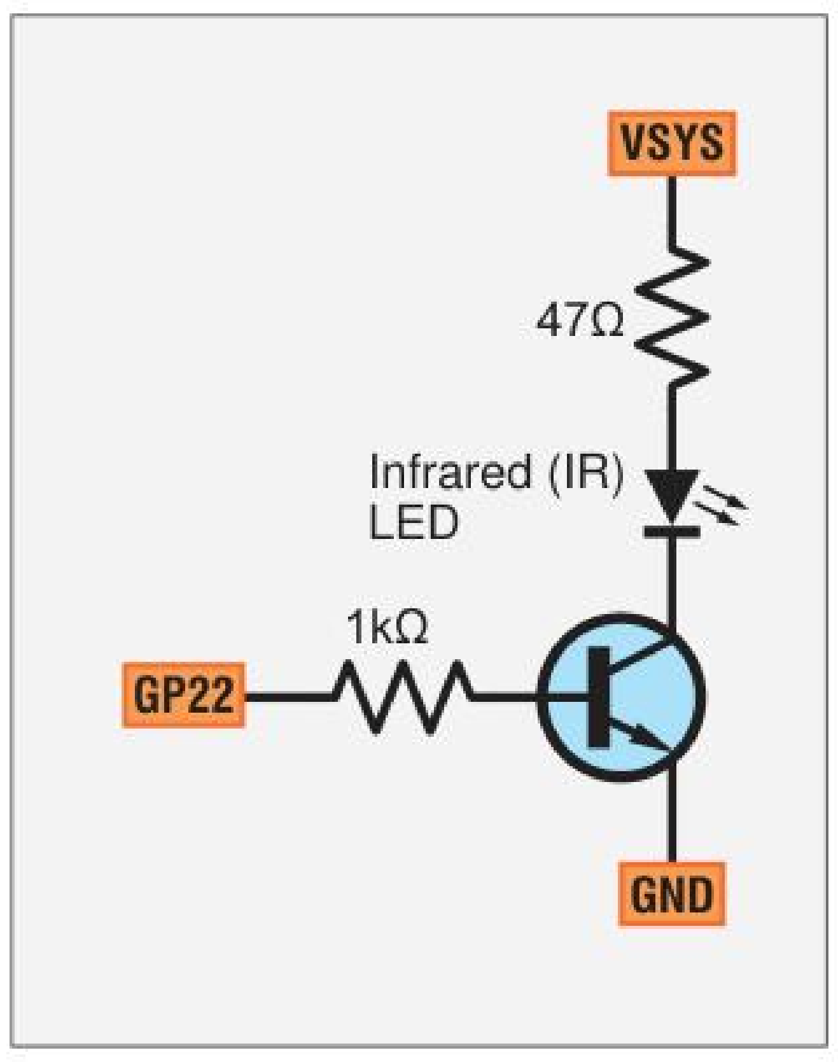


Fig.8. This simple IR transmitter circuit uses an NPN transistor, a couple of resistors, and an IR Tx LED.

code in this subroutine. Transmitting IR data is made simple thanks to the MMBASIC command `IR SEND` – please refer to the *PicoMite User Manual* for more details about the parameters used with this command.

Slave clocks

Each slave clock just needs a standard TSOP IR receiver connected to any spare GP input pin – in this case we will use GP22. Fig.10 shows the connections but note that we must connect the TSOP IR Rx to a 3.3V supply rather than the more usual 5V supply. The reason for this is that the maximum input voltage on any input pin (here GP22) is 3.6V. So, by supplying 3.3V to the TSOP, we will keep the input voltage on GP22 below the 3.6V limit and not damage the Pico.

Connect a TSOP IR receiver to GP22, GND and 3V3, as shown in Fig.10. We used three female-female DuPont leads with the TSOP inserted at one end, and the other end used to connect to a 1.3-inch PicoMite clock (that was moved onto an Expander module – see Fig.11). The program code can then use the standard IR interrupt (set up with the two commands: `SETPIN GP22, IR` and `IR DevCode, KeyCode, myIrInt`). This will result in the program code jumping to a subroutine (here called `myIrInt`) whenever an IR signal is detected. We then put the IR receiving code in this subroutine to process, check, and if necessary, update the `TIME$` variable.

Software

The two required program listings are a bit too long to print line-for-line here, so we have made two downloads available at the September 2022 page of the *PE* website. Download the **MasterClockIR.txt** and **SlaveClockIR.txt** files and load them into the relevant PicoMite clocks. `RUN` both programs, and if using a PICO-LCD-0.96 as the master clock, press the 'B' button to sync and update any slave clock that is within range of the master clock. If you are not using a PICO-LCD-0.96 as the master, you will need to modify the start of the MasterClockIR program to allow for a suitable trigger (see comments within the code for how to do this).

Epoch time

If you take a closer look at the program code, you may spot two new commands: `EPOCH(now)` and `DATETIME$(`*`n`*`)`. The first command converts the current date/time into 'Epoch time' (a number which represents how many seconds have elapsed since midnight GMT on 1 January 1970). The second command converts an Epoch time value into the usual date/time format. Why use these? Epoch time makes it much easier to do date/time calculations. For example, to add an hour onto the current time, convert the current time into Epoch time, add 3600 and convert back to normal date/time. This way, we don't have to worry about roll-over for something awkward like 31 December 1999 at 23:34:56 (which would mean rolling over date, month and year, as well as the hour to '00') all very tricky when you start looking into it – but very easy with conversion to an Epoch-time value.

Further examination of the MasterClockIR code will reveal that it is actually the Epoch time that is transmitted via IR, and the SlaveClockIR code converts it back into a recognised date/time. We do this because it means we only need to transmit five 7-bit key code values to transfer both date and time (refer to code comments for more details).

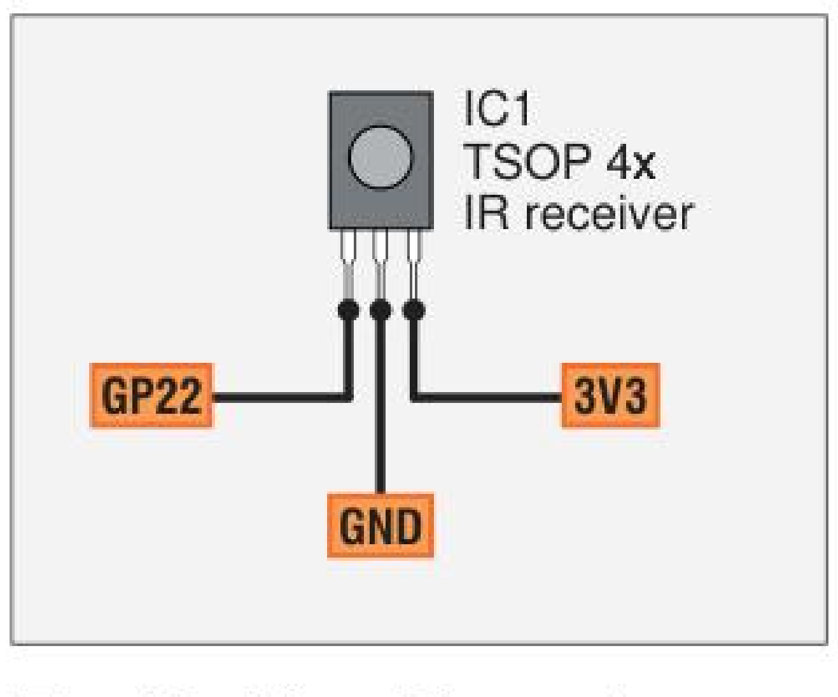


Fig.10. The IR receiver comprises just a single component – a TSOP IR Rx. Ensure it is connected to the 3.3V supply – not 5V.

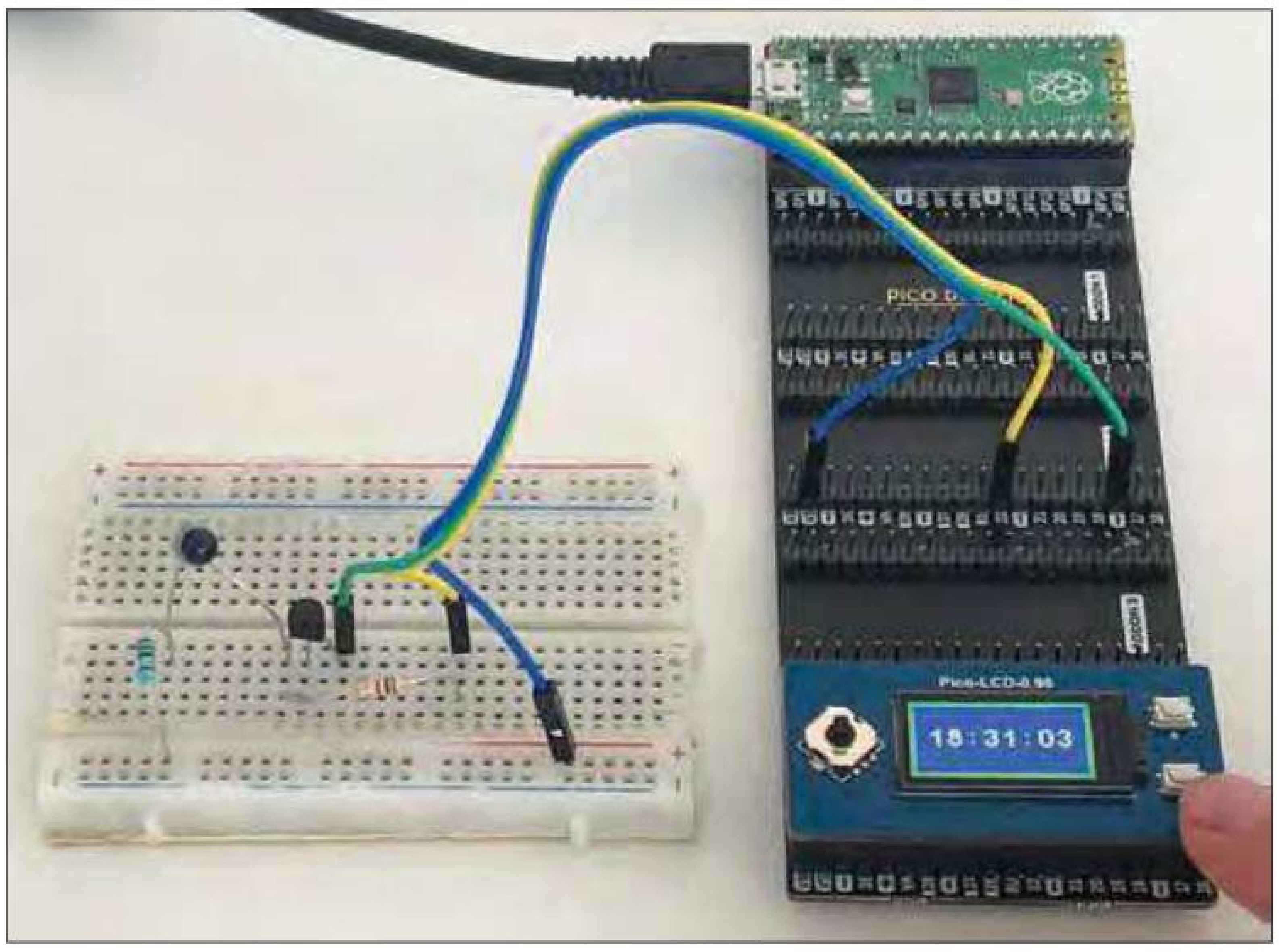


Fig.9. A Pico-LCD-0.96 master clock built on an expander module allows easy connection via three DuPont leads to the IR Tx circuit (which here has been assembled on a breadboard). Button B is used to send the time data via IR.

I hope this has shown you how easy it is to transmit a few bytes of data between PicoMites. Why not have a think about how you can use IR in your project other than to decode button presses from an IR remote control. If you come up with anything interesting, then do drop us an email and maybe we can mention it in a future article.

Next time

Next month, we will continue our exploration of the PicoMite and show you how to use a GPS receiver (another low-cost Pico plug-in module) to create a simple and useful tracker.

Until then, stay safe, and have FUN!

Questions? Please email Phil at:
contactus@micromite.org

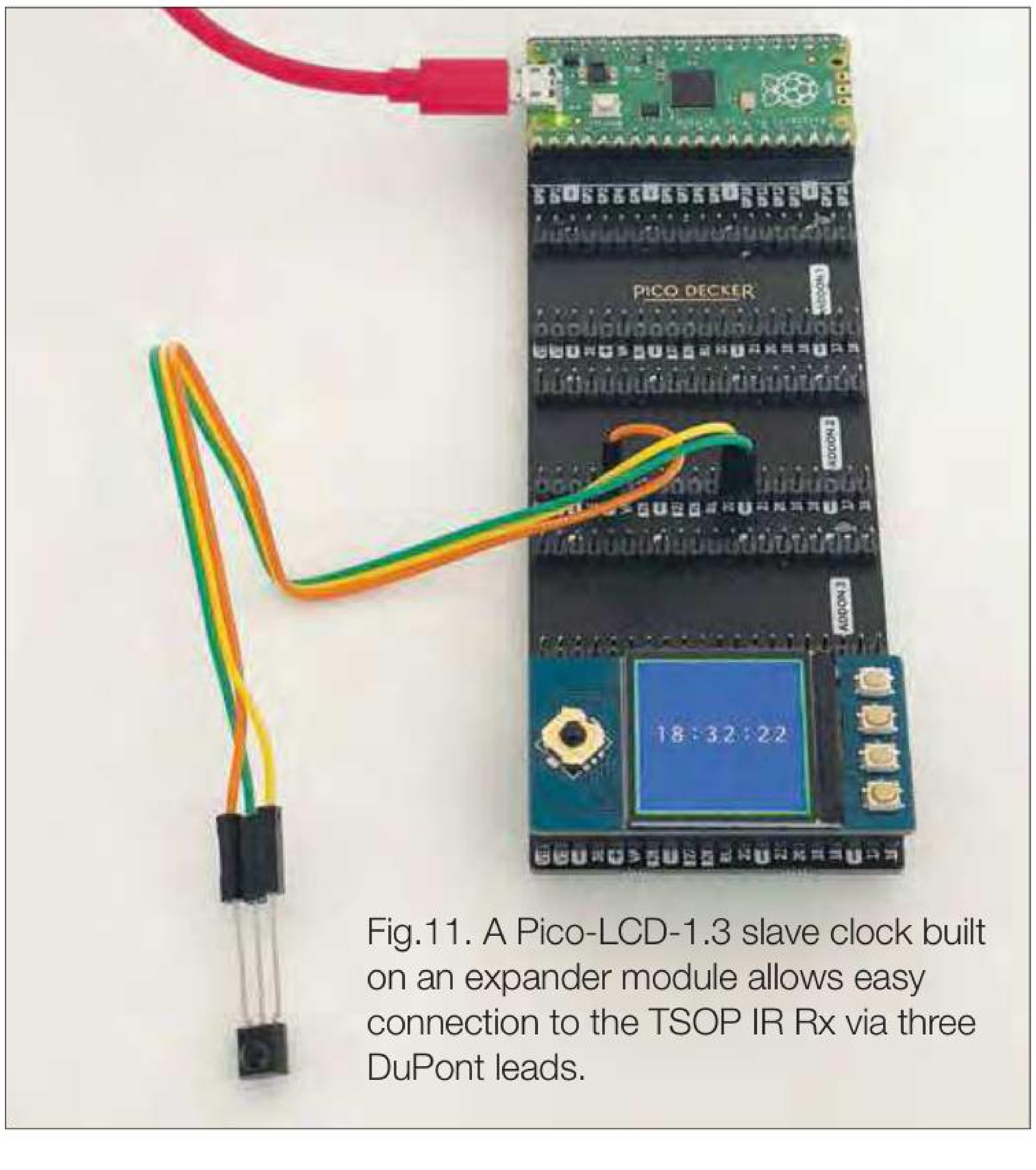


Fig.11. A Pico-LCD-1.3 slave clock built on an expander module allows easy connection to the TSOP IR Rx via three DuPont leads.

Max's Cool Beans

By Max the Magnificent

Flashing LEDs and drooling engineers – Part 31

As is usually the case, I'm currently bouncing off the walls with excitement. Sometime last year, I may have mentioned that I've been invited to give the keynote presentation at the FPGA Forum 2022 in Norway (**www.fpga-forum.no**). This auspicious affair is attended by all of the movers and shakers in Norwegian programmable space (where no one can hear you scream), including designers, project managers, technical managers, researchers, final-year students and FPGA vendors.

Norway Ho!

This year's Forum was originally intended to take place in February but... COVID... so it was rescheduled for September. The reason this is particularly poignant for me because I previously gave the keynote presentation at this exceptional event deep in the mists of time we used to call 2012. On the one hand, that's only 10 short years ago, which would be but a trifling hop in my time machine if I could ever get the little scamp working (you simply can't get the parts where I live in Huntsville, Alabama). On the other hand, technology has been racing along, as is its wont, and much has changed. For example, some FPGA companies that were with us in 2012 (eg, TierLogic and Tabula) have given up the ghost, bitten the dust, and shuffled off this mortal coil (I never metaphor I didn't like). Contrariwise, some feisty newcomers (eg, Efinix and Renesas) have decided to scamper over to the bar and join the party.

Oh yes, and there's also the fact that we are now up to our metaphorical armpits in artificial intelligence (AI), machine learning (ML) and deep learning (DL). Plus, we are being inundated with new flavours of reality (I'm a sucker for strawberry myself), including augmented reality (AR), diminished reality (DR), virtual reality (VR) and augmented virtuality (AV) to name but a few. Phew! Suffice it to say that I have little doubt I'll manage to scrape together something to talk about (much like my dear old mother, the real trick is to get us to *stop* talking).

Oftentimes, in addition to the Norwegian contingent, attendees to the Forum come from nearby nations, including Sweden and Finland. I'm also hoping to see representatives from a company called Testonica (**testonica.com**), which is headquartered in the Baltic country of Estonia. These guys and gals have a tasty technology called Quick Instruments. If you are designing a circuit board featuring an FPGA, all you have to do is provide a high-level view of the system – including the types of other devices on the board and their pin maps and register maps – and Quick Instruments automatically generates and compiles a corresponding firmware test suite, thereby allowing the FPGA to perform a sophisticated board-level self-test.

On the eve of a new era

The FPGA Forum has traditionally been held in the city of Trondheim, which is the third-most populous municipality in Norway. Among many other things, Trondheim is home to the Norwegian University of Science and Technology (NTNU). There are two reasons I mention this here. First, the day before my keynote, I'll be giving a guest lecture to a gaggle of MSc electrical and electronic engineering students at NTNU (**www.ntnu.edu**). Second, the company Halodi Robotics (**www.halodi.com**), which I introduced in my previous column (*PE*, August 2022), is headquartered in Norway and their EVE humanoid robots (Fig.1) can be found in most Norwegian universities. I'm hoping an EVE will deign to attend my lecture, in which case you can bet your cotton socks I'll be taking photographs and reporting further in a future column.

EVE is human-sized and has 23 degrees of freedom. Boasting two-arm manipulation, EVE can handle an 8kg payload with each arm. EVE can also squat down to pick things up off the floor or out of cupboards and reach up to place objects on shelves.

The amazing thing for me is that EVEs are already deployed and working around the world in applications ranging from retail to security. In the case of retail, for example, EVEs can perform tasks like roaming a supermarket's aisles, spotting out-of-place items, and returning them to their rightful locations, restocking shelves, gathering items for remote shoppers to pick up later, identifying potential hazards like spills and reporting them, and... the list goes on.

Similarly, when it comes to security applications, EVEs can autonomously roam around buildings looking for unauthorised intruders with unrecognised faces, checking doors are locked and lights are turned off… and so on and so forth. In fact, earlier this year, the American security company ADT signed the world's largest order to date for humanoid robots by ordering 140 EVEs from Halodi.

It all starts with a single servo

As amazing as EVE is, such creations all start with a single actuator like a servo, which is the point you and I currently find ourselves in our discussions. As discussed last month, we started with a 9g hobby micro servo, where the '9g' refers to the approximate weight of the servo, not the amount of torque it can provide.

Originally, I was planning on taking one of these bodacious beauties apart to show you what's inside, so I ordered a set

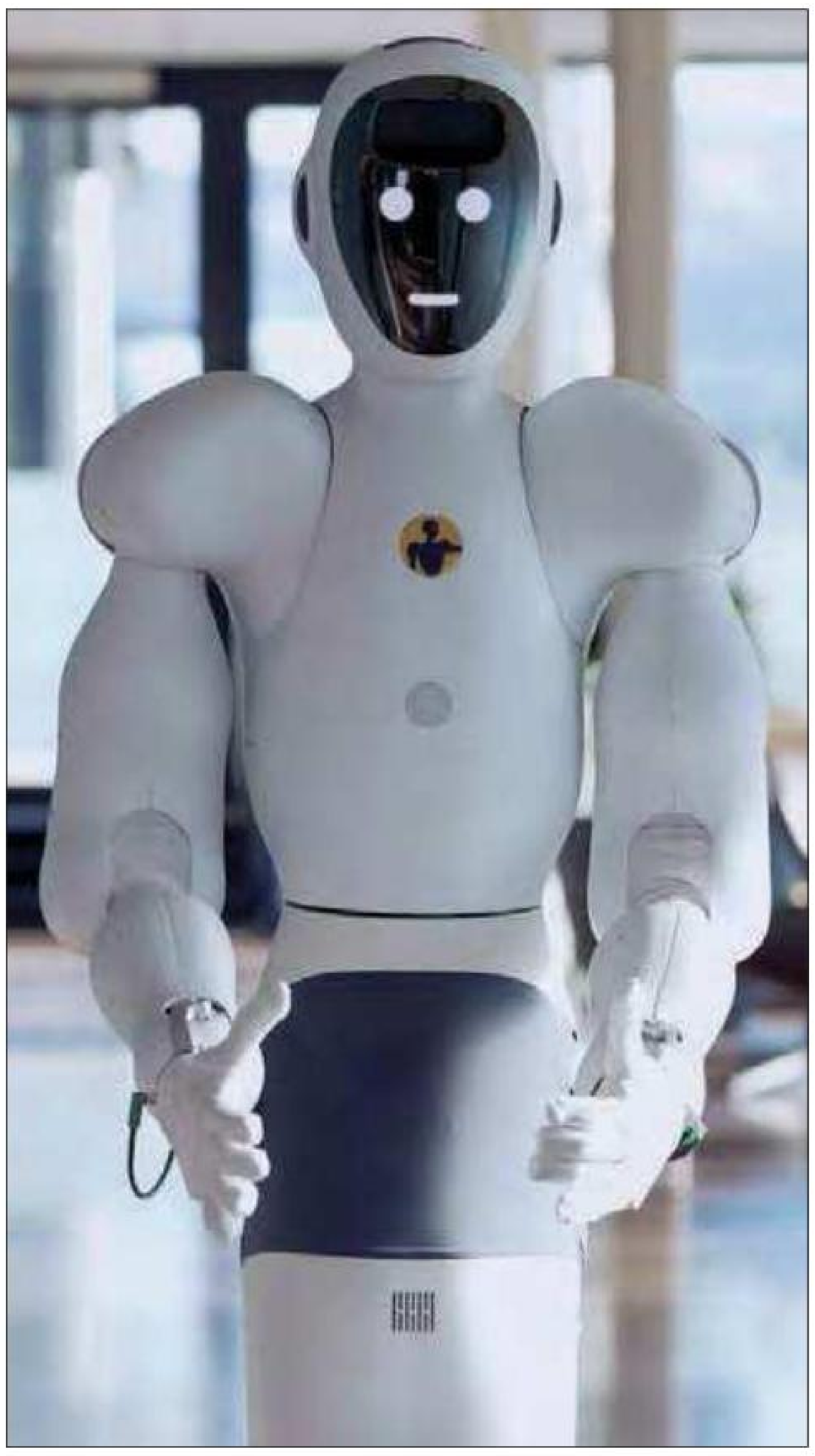


Fig.1. EVE the humanoid robot. (Image: Halodi Robotics)

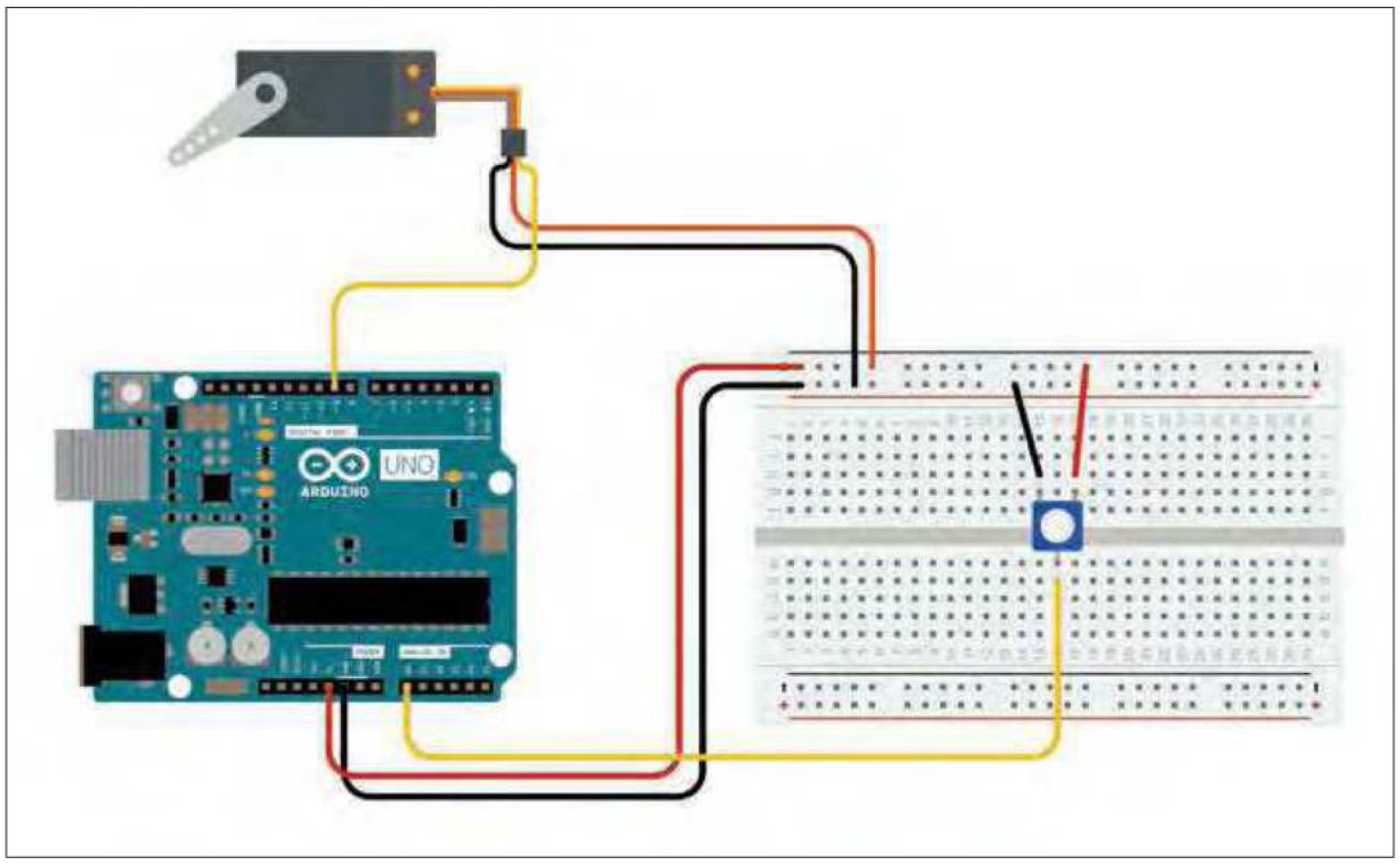


Fig.2. Knob Circuit in the theoretical world. (Image: Arduino)

of five SG90 micro servos from Amazon (I'll explain why my plans changed in a moment). There are myriad suppliers for this sort of thing, including many generic offerings (**https://amzn.to/3O32sC4**), but 'you get what you pay for,' so be careful what you order.

The cheap-and-cheerful servos I opted for offer 180° of rotational motion. As we previously discussed, the way these servos work is that our controller – an Arduino Uno is this case – periodically sends a positive-going pulse to the servo. The period between pulses isn't particularly important – it's the width of the pulses that counts – but it's common to use a 'refresh' frequency of 20ms (milliseconds), which means we send new pulses to the servo 50 times each second (50Hz).

When it comes to the width of the pulses, a value of 1.5ms instructs the servo to move to its default (centre) position. In the case of the servos I ordered, the accompanying data sheet says that a pulse of 1.0ms corresponds to a rotation of –90° (anticlockwise all the way to the 'left') while a pulse of 2.0ms corresponds to a rotation of +90° (clockwise all the way to the 'right').

Although a 1.5ms pulse will always direct this sort of servo to its centre position, different servos may vary with respect to their end-point values. According to *Motors for Makers* (**https://amzn.to/3IovsD0**), for example, some servos will accept pulses from 0.7ms (full rotation left/anticlockwise) to 2.3ms (full rotation right/clockwise) (by 'full' we mean the maximum rotation supported by that particular servo).

The main thing is that we don't want to make our pulses too narrow or two wide such that they exceed the servo's range because this may result in our stripping the gears, which is not considered to be a good thing to happen.

It's also important to remember that these servos are intended for hobby applications, so their specifications and tolerances can be more than a little wibbly-wobbly compared to industrial-grade devices. What I'm trying to say is that, even in a batch of supposedly 'identical' servos, there is the potential for a lot of variation, so it's a good idea to characterise each device before using it. Speaking of which...

Turn the knob on your side

Older readers may remember a British radio comedy called *The Goon Show* featuring Spike Milligan, Harry Secombe, Peter Sellers and Michael Bentine. This zany program, which was a favourite of Prince Charles, was originally broadcast from 1951 to 1960 with occasional repeats. I recall an episode during which one of the characters said, 'Turn the knob on your side.' Another character responded, 'I haven't got a knob on my side.' And the first character retorted, 'Of the door, you idiot!' (You had to be there and hear their voices. I'm still chuckling inside.)

In my previous column, we noted that the Arduino comes equipped with a servo library (**https://bit.ly/3O3rIZu**). We also created a 'Sweep' circuit and program that caused our servo to sweep back and forth. Now we're going to implement what the folks at Arduino call their 'Knob' circuit.

It's pleasing how pristine things are in the theoretical world (Fig.2). It's also remarkable how much 'rattier' they appear when viewed in the real world (Fig.3). In this case, as opposed to the breadboard-mounted potentiometer (pot) shown in the theoretical circuit, I'm using a rather chunky free-standing pot as shown in the upper right-hand corner of the image.

Now let's consider the program we're going to use to characterise our servo (Fig.4). Our first step on Line 1 is to include the servo library. On Line 3, we create a servo object. In this case, we've called it `MyServo`, but any legitimate (non-keyword) name will do. On Line 5 we declare an integer (`int`) variable called `PotPin` that we assign to the Arduino's `A0` analog input, and then on Line 6 we declare another integer variable called `PotVal` that we will use to store the values we read from our pot.

On Line 10, we use the `attach()` method to attach our servo variable to a pin. In this example we employ three arguments: 9, 1000 and 2000. The first argument, which is mandatory, specifies the pin we wish to use to drive our servo. This must be one of the pins that support pulse-width modulation (PWM), which are pins 3, 5, 6, 9, 10 and 11 in the case of an Arduino Uno (we are using pin 9 in this example). The second and third arguments, which are in units of microseconds, are optional. The second argument specifies the pulse width corresponding to the minimum (0°) angle on the servo, while the third argument specifies the pulse width corresponding to the maximum (180°) angle on the servo. We are using values of 1,000µs and 2,000µs (1ms and 2ms), respectively. Interestingly, the default values used for these parameters if we don't specify them are 544 and 2400, respectively, which are well outside the range values we have been led to regard as being valid.

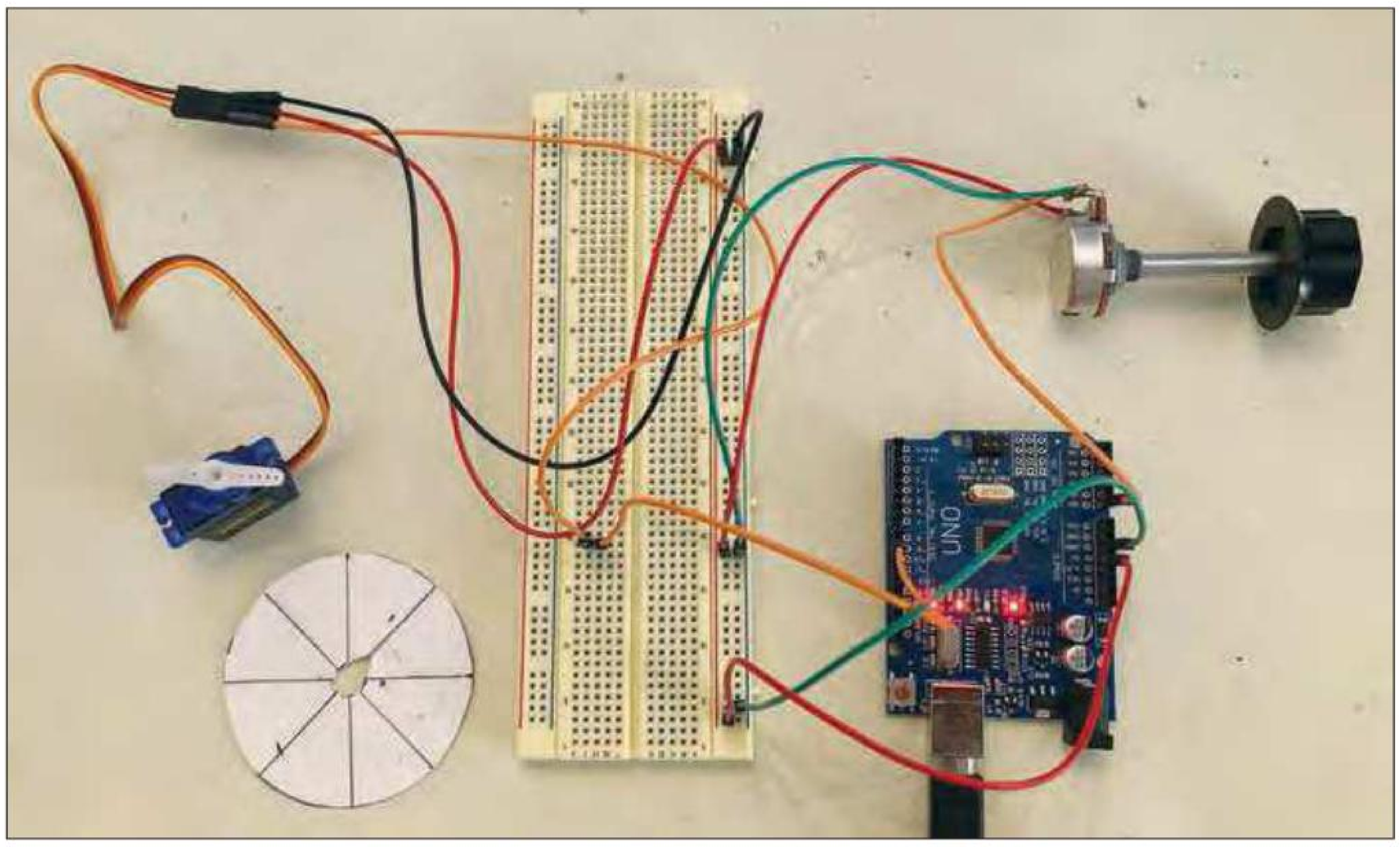
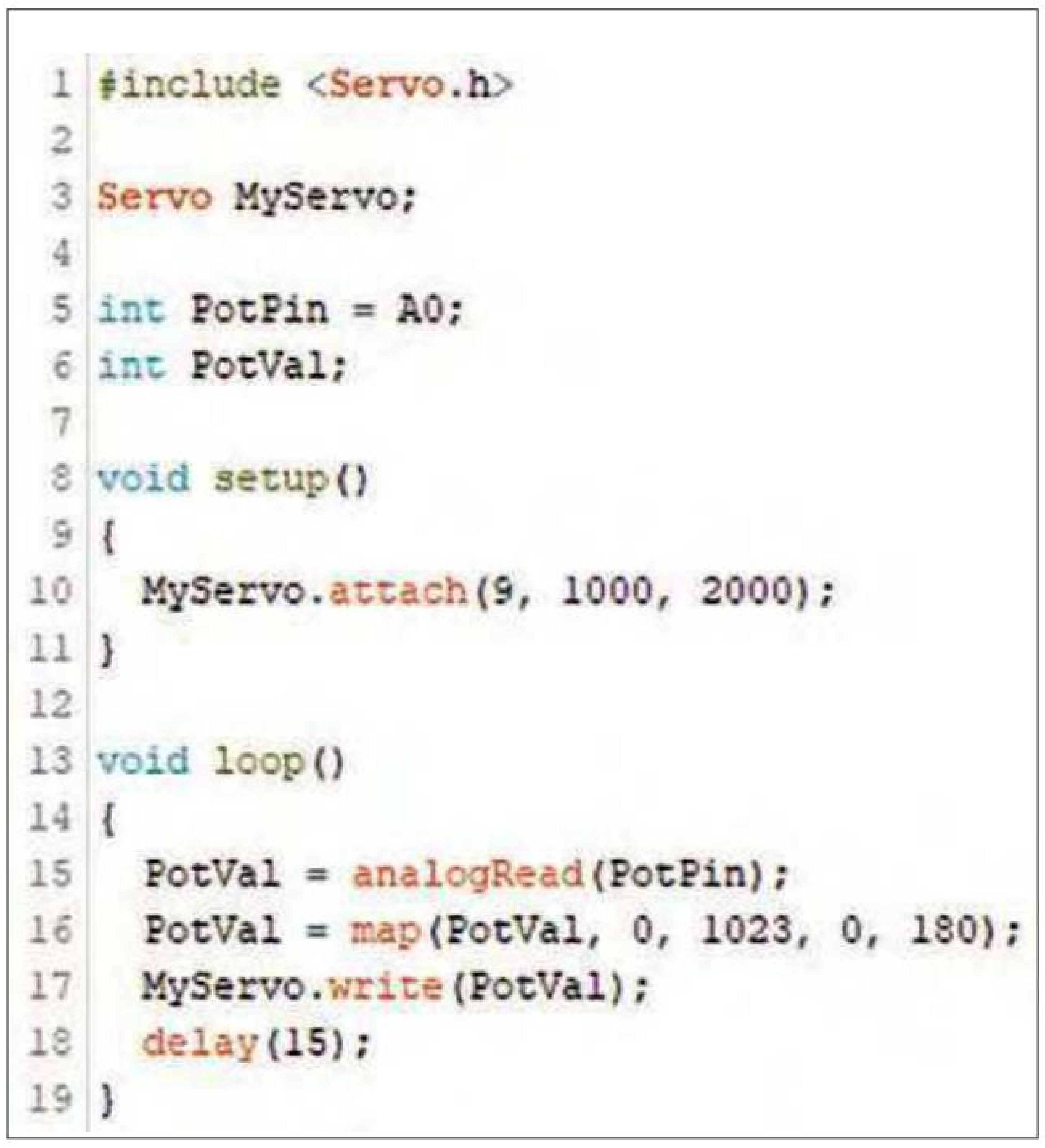


Fig.3. Knob Circuit in the real world (the blue servo appears on the left).



```
1  #include <Servo.h>
2
3  Servo MyServo;
4
5  int PotPin = A0;
6  int PotVal;
7
8  void setup()
9  {
10   MyServo.attach(9, 1000, 2000);
11 }
12
13 void loop()
14 {
15   PotVal = analogRead(PotPin);
16   PotVal = map(PotVal, 0, 1023, 0, 180);
17   MyServo.write(PotVal);
18   delay(15);
19 }
```

Fig.4. Example Knob Circuit program.

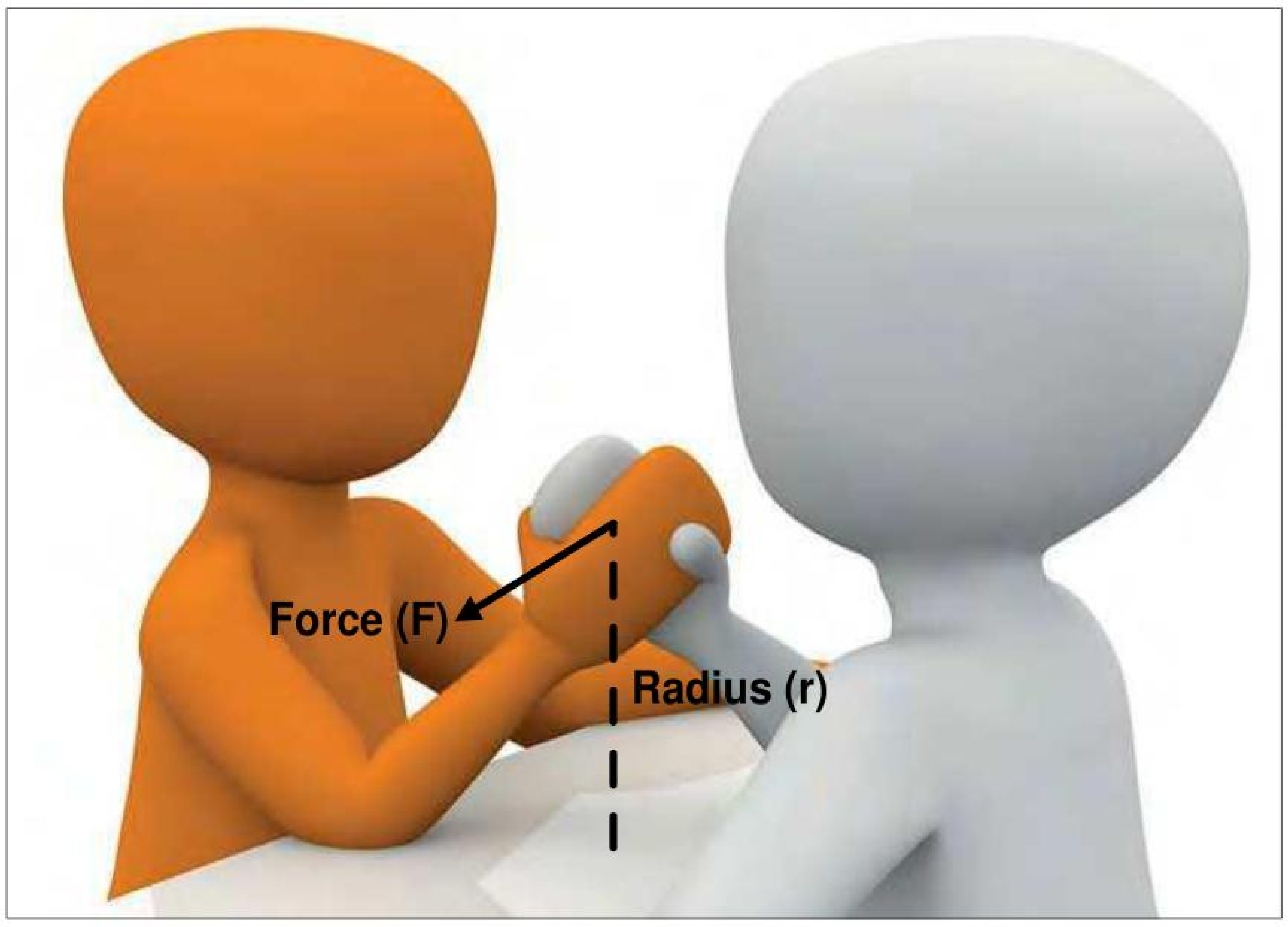


Fig.5. Visualising torque.

When we come to the main loop, we start on Line 15 by reading the value from the pot into our `PotVal` variable. Since the Arduino Uno has a 10-bit ADC (analogue-to-digital converter), this will result in values between 0 and 1023. The `write()` method provided by the Arduino's servo library requires values in the range 0 to 180, corresponding to 0° to 180°, respectively. Thus, on Line 16, we use the Arduino's `map()` function to convert the 0 to 1023 values read from the pot into their 0 to 180 equivalents. On Line 17, we write this new value to the servo, and on Line 18 we insert a 15ms delay to give the servo time to respond.

A puzzling poser

In Fig.3 you'll see a small circle of card marked out in 45° angles in the lower left of the image. This is just something I threw together for the purposes of these discussions. If I had been doing this in earnest, I would have created a more accurate tool divided into 5° segments.

When I placed this card under the rotating arm on top of the servo and ran the program above, turning the pot from one extreme to the other caused the servo arm to move only 90°. I changed the `attach()` method's min and max arguments to be 700 and 2300, respectively. This resulted in an approximately 135° swing of the servo arm. Finally, I tried the servo library's default values of 544 and 2400, which resulted in something close to a 180° swing, although I did detect little 'graunching' sounds at the extremities of the movement.

This left me perplexed and puzzled. I can accept a little variance, but why would the data sheet specify min and max values of 1000 and 2000 when the servo actually requires values of 544 and 2500? (If you have any thoughts on this, I'd love to hear them.)

Not so simple

I fear this is the point in this article where we may expect some gnashing of teeth and rending of garb because our illustrious publisher, Matt Pulzer, keeps on telling me to 'stick to the electronics' and I keep on saying to myself, 'But this other stuff is soooo interesting!'

For example... the term 'simple machine' refers to a mechanical device that can be used to change the direction or magnitude of a force. The six classical simple, as defined by Renaissance thinkers, are the lever, wheel and axle, pully, inclined plane, wedge and screw.

I could waffle on about these little beauties for ages, but we have other fish to fry. Of particular interest to us here is the fact that gears may be considered simple machines that consist of two or more wheels equipped with interlocking teeth such that when one gear is rotated any other interlocked gears rotate in the opposite direction.

It's probably worth noting that some people – especially in the UK – informally refer to gears (aka gearwheels or cogwheels) as cogs, but the term 'cog' officially denotes an individual tooth on the gear.

Although they may appear simple, there's much more to gears than meets the eye. We could spend weeks delving into their intricacies, but we won't because I can feel Matt's all-seeing eye starting to swing in my direction.

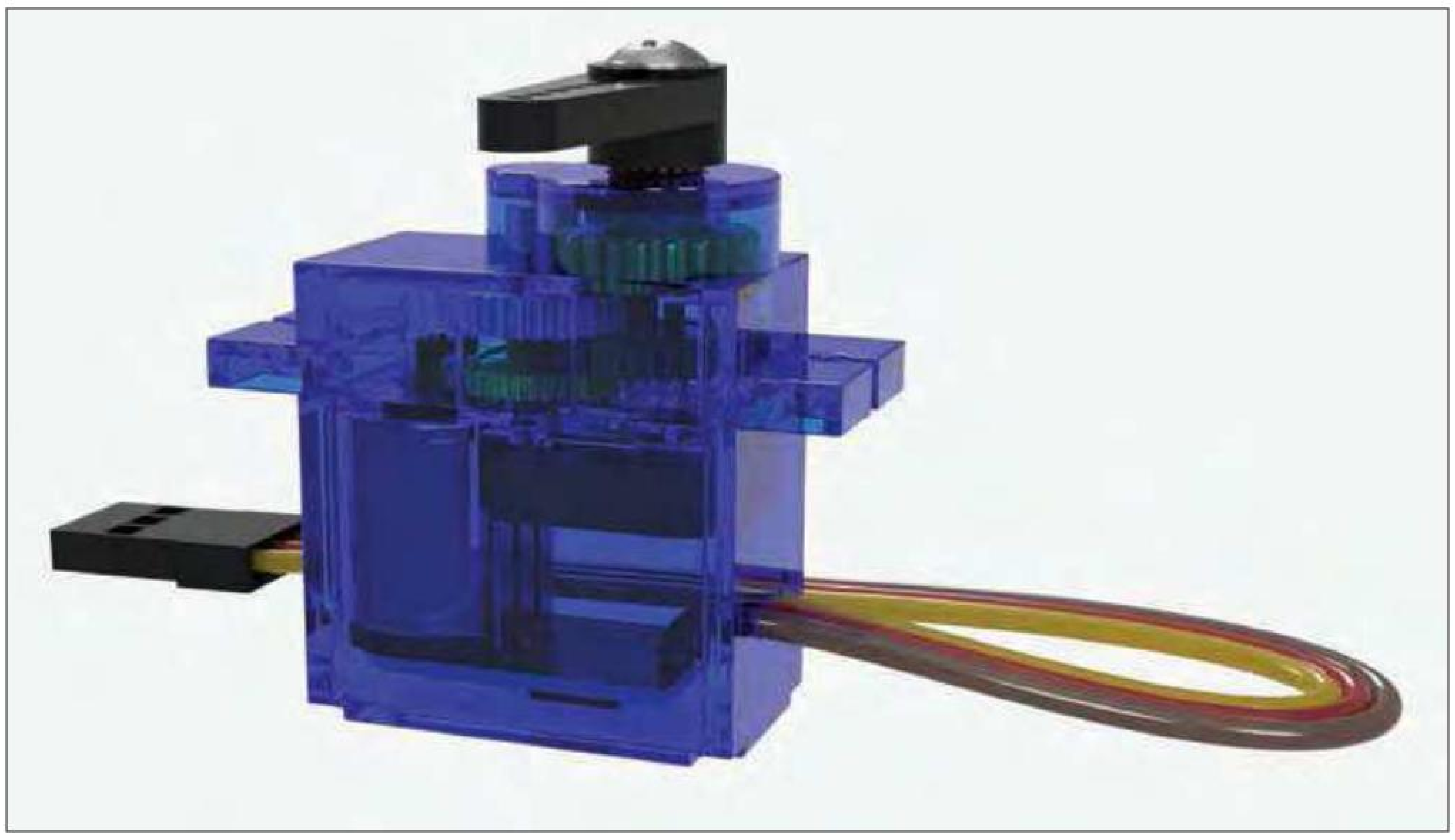


Fig.6. Classic servomotor with arm attached to the output of the gear train.

You can torque!

I'm afraid this is the point where we are going to have to introduce some semi-technical topics. Let's start with torque, symbol T, which is the rotational equivalent of linear force and which we might think of as a 'twisting force.' The easiest way to visualise this is as an arm-wrestling competition (Fig.5).

If we assume both contenders are equally matched, then the amount of force they are applying will cancel each other out, but that's not the point we're trying to make. If we consider only the nearside character, the amount of torque equals the force multiplied by the radius: $T = r \times F$.

There are two ways to look at this. First, let's suppose you are trying to open a heavy hinged door by pushing on it. In this case,

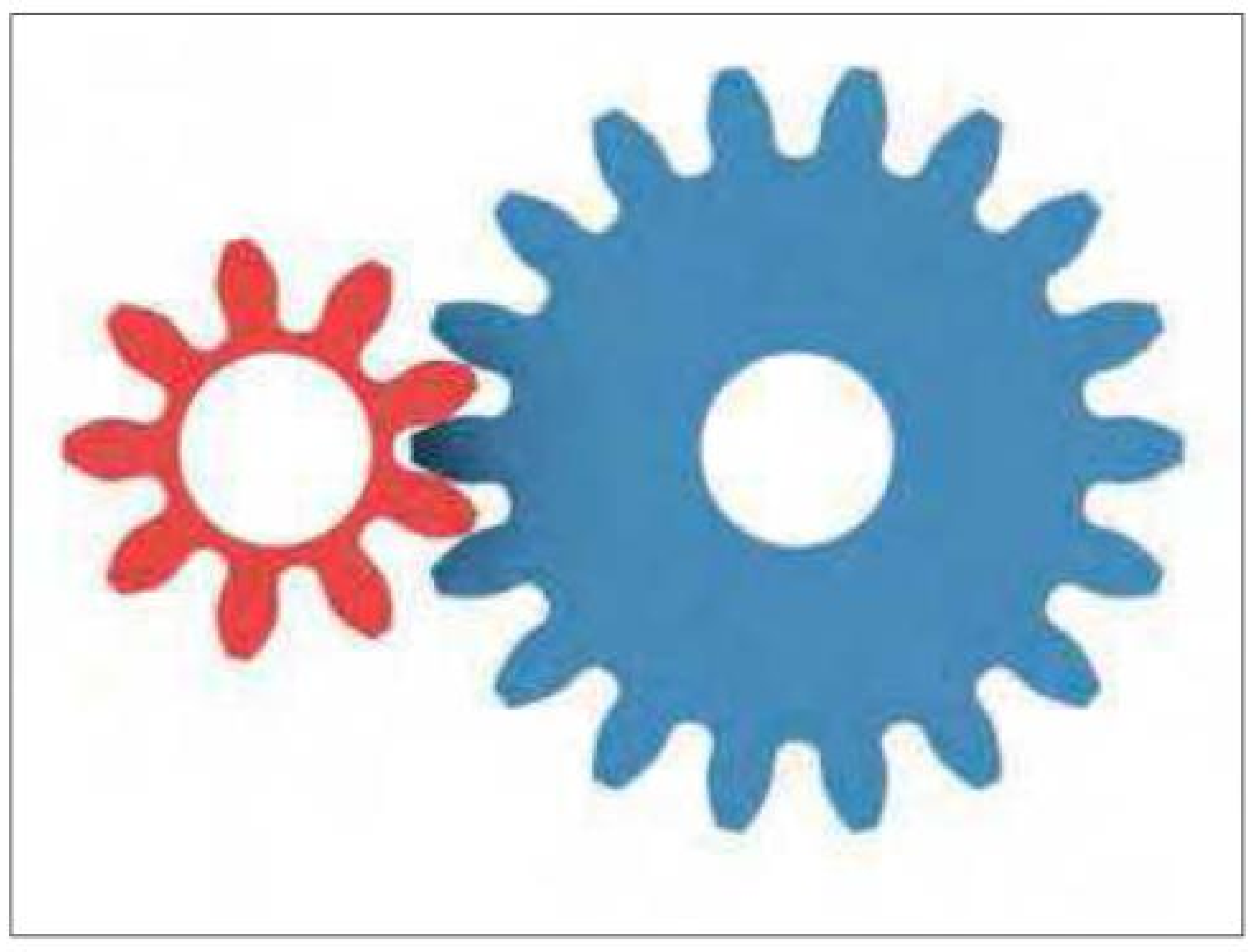


Fig.7. Two gears with 9 and 18 teeth. (Image: Steve Manley)

the maximum amount of force you can apply (ie, how hard you can push) is fixed. The radius is the distance from the hinges to the point at which you are pushing. If you push close to the hinges, the radius is small, so the torque you are generating will be small, which means opening the door will be difficult. By comparison, if you push on the door as far away from the hinges as you can, then the radius is larger, which means that – with the same amount of force as before – you can generate more torque, thereby (a) making opening the door a breeze and (b) letting a breeze pass through the door.

The second way to look at things might be from the point of a servo with an arm on its output (Fig.6). In this case, the amount of torque that can be generated by the motor is fixed, so the amount of force the servo can exert is given by $F = T/r$. This tells us that the greater the radius of the arm (at the point we connect our linkage to it), the less force it can exert.

Another concept is that of angular speed, symbol ω, which describes the angle an object moves through in a given amount of time. In the case of electric motors, angles are measured in degrees (symbol °), and angular speed is measured in revolutions per minute (RPM).

When two gears are interlocked, the gear with fewer teeth must rotate more times than the gear with more teeth. Consider two gears where one has 9 teeth and the other has 18 teeth (Fig.7). In this case, the smaller gear will rotate two times for every rotation of the larger gear.

What happens next depends on which of the gears is the driving gear (the input) and which of the gears is being driven (the output). If the larger gear is the input, then – in the case of our 9- and 18-tooth example – the smaller output gear will rotate at twice the speed but with only half the torque (twisting force). By comparison, if the smaller gear is the input – which is the more common case (and which is the way things work in our servo) – then the larger output gear will rotate at only half the speed but with twice the torque.

The term 'gear ratio' refers to the relative torque between two gearwheels and is typically expressed as X:1, where X is the proportional increase in torque. In the case of our example system, the gear ratio would be 2:1. If our larger gear had 27 teeth then the gear ratio would be 3:1, and so on. When two or more gears are connected, we call the result a 'gear train.' If a 2:1 gear train is connected to a 3:1 gear train, which is – in-turn – connected to a 4:1 gear train, for example, then the resulting gear ratio for the entire gear train will be $(2 \times 3 \times 4){:}1 = 24{:}1$.

Gears are typically very efficient, with power losses due to friction accounting for only a few percent. To make our lives simple for the purposes of these discussions, we will assume 100% efficiency.

Just to round out this part of our discussions, in the context of gears, power (P) equates to torque multiplied by rotational speed: $P = T \times \omega$. For our 9-pin input and 18-pin output gears, $P_I = T_I \times \omega_I$ and $P_O = T_O \times \omega_O$, respectively. Assuming 100% efficiency, $P_O = P_I$, which means $T_O \times \omega_O = T_I \times \omega_I$. In turn, this means $T_O/T_I = \omega_I/\omega_O$.

Furthermore, if we use N to denote the number of teeth on a gear wheel, then the relationship between angular speed and the number of teeth can be expressed as $\omega_I/\omega_O = N_O/N_I$. Tying all of this together, we can say our gear ratio of $2{:}1 = T_O/T_I = \omega_I/\omega_O = N_O/N_I$. (Note the similarity to a transformer's turns ratio, voltages and currents.)

Feast your orbs

As I mentioned earlier, I originally planned on dismantling one of my cheap-and-cheerful servos and taking pictures to show you what was inside. However, my friend Steve Manley said that he could do a much better job using his Fusion 360 computer-aided design (CAD) software.

Steve started by purchasing a couple of AZ-Delivery MG90S micro servos from Amazon (**https://amzn.to/3ywy4Kk**). He then dismantled one, measured everything with his micrometer, and used his microscope to painstakingly count the teeth on each of the gears forming the gear train. The result was a set of images that brought tears of joy to my eyes (Fig.8).

One thing Steve noted is that he was led to believe that all of the gears in this servo were metal. However, it turned out that only the output gear was metal while the rest were plastic. Looking at the comments on Amazon, several other people were disgruntled by this deception. (As always, *caveat emptor* is the order of the day.)

The term 'compound gear' refers to two or more gears that are fixed together and therefore rotate at the same speed. The green, blue, and red gears in Fig.7 are compound gears because they are each composed of two gears.

In the case of this servo, we have three shafts: the motor shaft, an intermediate shaft, and the output shaft. It's important to note that the green and blue compound gears are not physically attached to the intermediate shaft; they just rotate around it. Similarly, the red compound gear is not physically attached to the output shaft. The only gears that are physically attached to their associated shafts are the brass-coloured input gear that's attached to the motor shaft and the gold-coloured output gear that's attached to the output shaft.

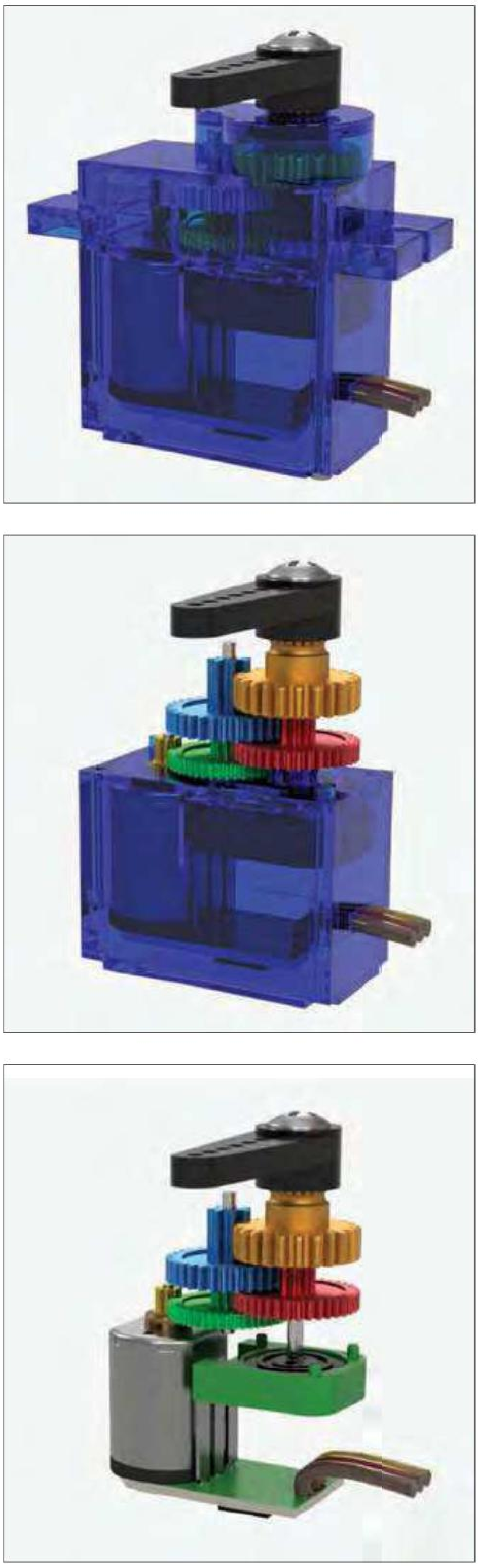


Fig.8. Servo in case (top) and with case removed (bottom). (Image: Steve Manley)

Observe the black arm on the top of the servo. This may also be referred to as the 'horn.' These horns come in a variety of shapes and sizes. Also observe that the diameters of the shafts increase from input to output (motor shaft = 1mm, intermediate shaft = 1.16mm, output shaft = 1.36mm). Similarly, the size of the teeth and the thickness of the gears increases as we progress

	Shaft	Gear	# Teeth	Ratio
	Input	Input	9	
	Intermediate	1A	47	5.2222
		1B	10	
	Output	2A	38	3.8000
		2B	8	
	Intermediate	3A	32	4.0000
		3B	7	
	Output	Output	23	3.2857

Fig.9. Summary of servo gear train.

through the gear train. All of these changes are required to accommodate the increasing torque from input to output.

The green element located below the red compound gear contains a potentiometer that's attached to the motor shaft. This pot is used to measure the current angular position of the shaft (various other techniques may be employed by different servos).

The small printed circuit board (PCB) carrying the control electronics is mounted at the bottom of the servo under the motor. In addition to a voltage regulator and some discrete components, this board features a KC9702A control chip that interprets the PWM signal coming from the outside world, uses the value from the potentiometer to determine the difference between the desired and actual positions of the output shaft, and drives the DC motor accordingly.

What? You're joking!

A summary of the gears and gear ratios forming this servo's gear train is presented in Fig.9. If we multiply all of the gear ratios together (5.2222 × 3.8000 × 4.0000 × 3.2857, we end up with a total input to output drive train ratio of ~260:1.

This may not, at first glance, leave you quivering in excitement. After just a little contemplation, however, I think you'll agree the results are more than impressive.

Let's start with the fact that the torque of the motor itself is miniscule. For example, Steve notes that he can easily stall the motor by pinching its rotating shaft between his thumb and forefinger.

Steve is in the process of creating a video featuring an animation of this servo in action. I'm not sure from whence he obtained these numbers but, in his script, he starts by assuming that the motor's speed is 26,000 RPM and its torque is 0.0072 kg-cm. Also, the rotational speed (ω) of the horn, as specified in in the data sheet is 60°/0.1s. So, ω = 60°/0.1s = 360°/0.6s

= 1 rotation / 0.6s

= 100 rotations / 60s

= 100 RPM.

Assuming our gear ratio is 260:1 as determined above, then our speed reduction from input to output will be 26,000 RPM / 260 = 100 RPM, which matches the data sheet. Meanwhile, our torque increase will be 0.0072 kg-cm × 260, which results in 1.87 kg-cm torque at the servo's horn.

Now, if we ignore any 'knobbly' bits sticking out of the top and sides, using this gear train allows our servo to fit in a teeny-tiny package only around 23mm wide by 29mm tall by 12 mm deep. Suppose we were to return to the 2-gear scenario with which we started out in Fig.7. Further, let's assume we wish to keep our original motor with its 9-tooth gear, and we also want to maintain our gear ratio of 260:1. This means our output gear will need 9 × 260 = 2,340 teeth. In turn, this means the diameter of the output gear will be 468mm (Fig.10.).

I think we can all agree that having a gear close to half a meter in diameter would not be well-suited to small RC model applications. I don't know about you, but I find it amazing to think that, by means of a handful of modestly sized gears, we can reduce a half-meter monster down to a package of around eight cubic centimetres.

Next time

I'm afraid the time has come for me convey this column to head office and face 'The wrath of Matt' (I've just dispatched the butler to retrieve my 'quaking in fear' trousers). In my next column, in addition to being able to show Steve's animation demonstrating the internal workings of the servo in awesome detail, we will move on to using joysticks to control multiple servos at the same time. Until that propitious day, as always, I welcome your captivating comments, insightful questions, and sagacious suggestions.

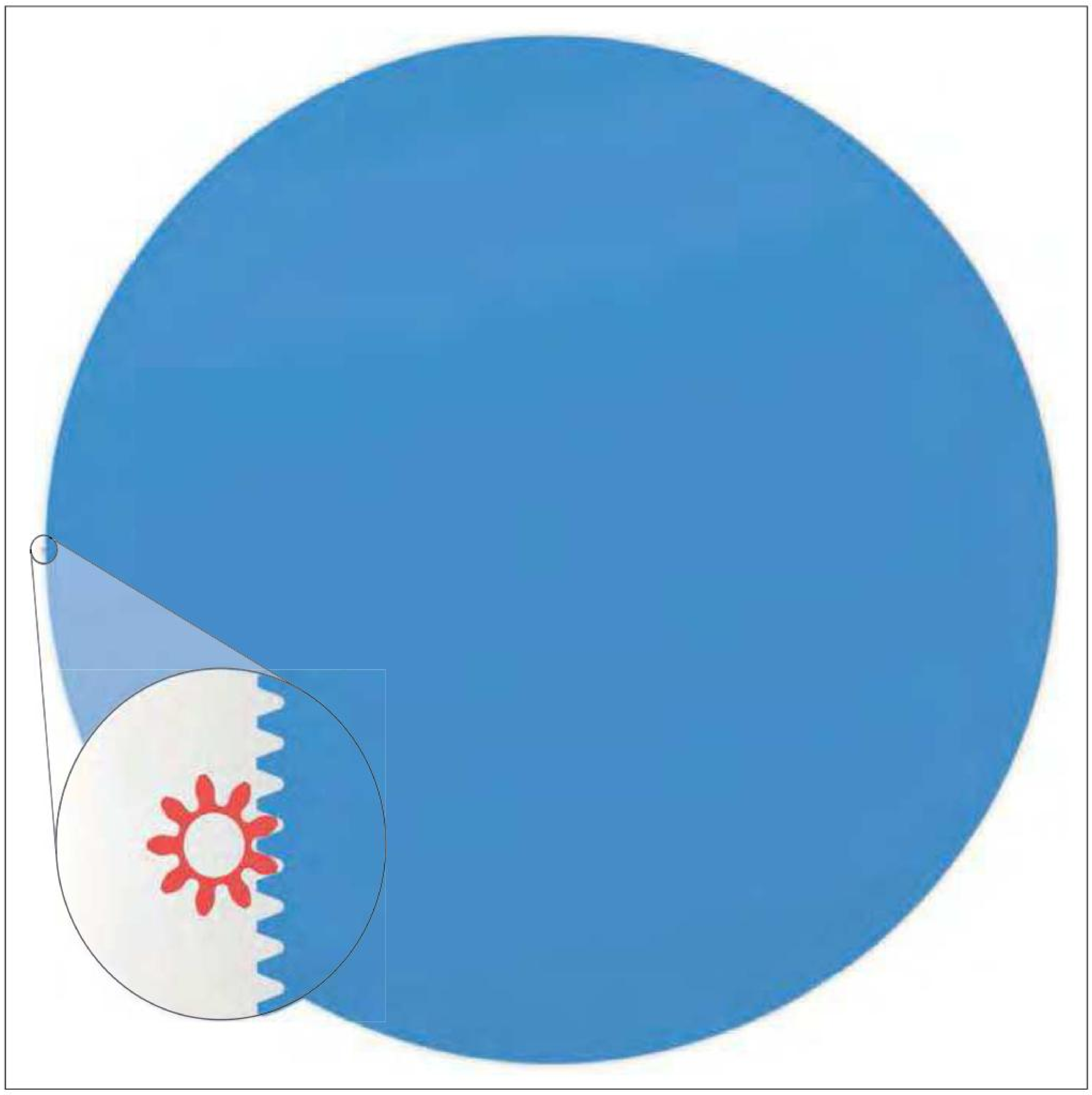


Fig.10. Without gear trains, hobby models would be bigger than the real thing.

Cool bean Max Maxfield (Hawaiian shirt, on the right) is emperor of all he surveys at **CliveMaxfield.com** – the go-to site for the latest and greatest in technological geekdom.

Comments or questions? Email Max at: **max@CliveMaxfield.com**

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Lots of big valve resistors available.

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AUDIO OUT

By Jake Rothman

Transformers in audio – Part 2

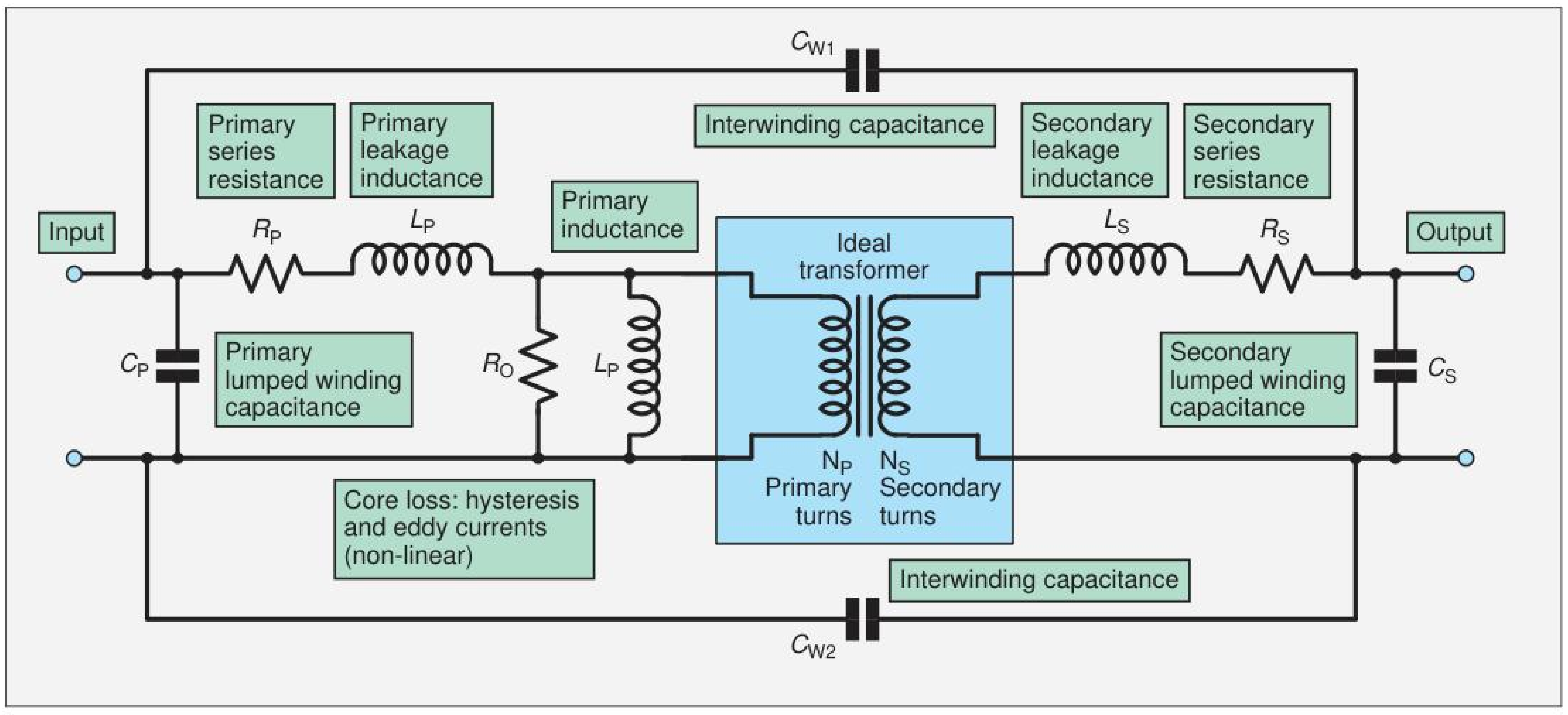

C_{W1}
Interwinding capacitance
Primary series resistance
Primary leakage inductance
Primary inductance
Secondary leakage inductance
Secondary series resistance
Input
R_P
L_P
Ideal transformer
L_S
R_S
Output
C_P
Primary lumped winding capacitance
R_O
L_P
Secondary lumped winding capacitance
C_S
N_P Primary turns
N_S Secondary turns
Core loss: hysteresis and eddy currents (non-linear)
Interwinding capacitance
C_{W2}

Fig.9. The equivalent circuit of a transformer with all the parasitic elements added.

Last time (July 2022) we looked at the uses and advantages of one of the oldest components in electronics – the transformer. This month, we will examine the limitations of real, as opposed to ideal transformers, as well as exploring some common audio transformer applications.

Transformer equivalent circuit and 'parasitics'

The theoretically ideal transformer has no losses and an infinitely wide frequency response. However, no component is perfect, although power transformers come close in terms of efficiency. Audio transformers have very real frequency response limitations at each end of the audio spectrum and because of the way induction operates – depending on a *rate of change* of magnetic flux – transformers cannot pass DC, which is generally a good thing in audio. The limitations of a real-world transformer lead us to the transformer equivalent circuit, which takes account of the non-ideal performance of real (as opposed to ideal) transformers.

Losses

All components can be represented by an equivalent circuit consisting of the ideal component with associated undesirable losses and 'parasitics,' which are represented by additional 'components'. This is shown in Fig.9 where additional passive components (resistors, capacitors and inductors) have been added to a truly ideal transformer.

The most obvious loss is the DC resistance of the windings, which are called 'copper losses' and are represented as resistances in series with the windings. Next, we have iron losses (core losses) due to eddy currents and hysteresis. Energy is lost by the emission of stray magnetic fields, which is worse for standard transformers compared to toroidal transformers. These losses are real. The old crude doorbell transformer in my house continuously dissipated a significant 2W of heat. Replacing it with a toroid eliminated this energy loss. Over 22 years I paid £85 for a lot of electricity units (385kWh) to run a doorbell that only rang for around three seconds a day!

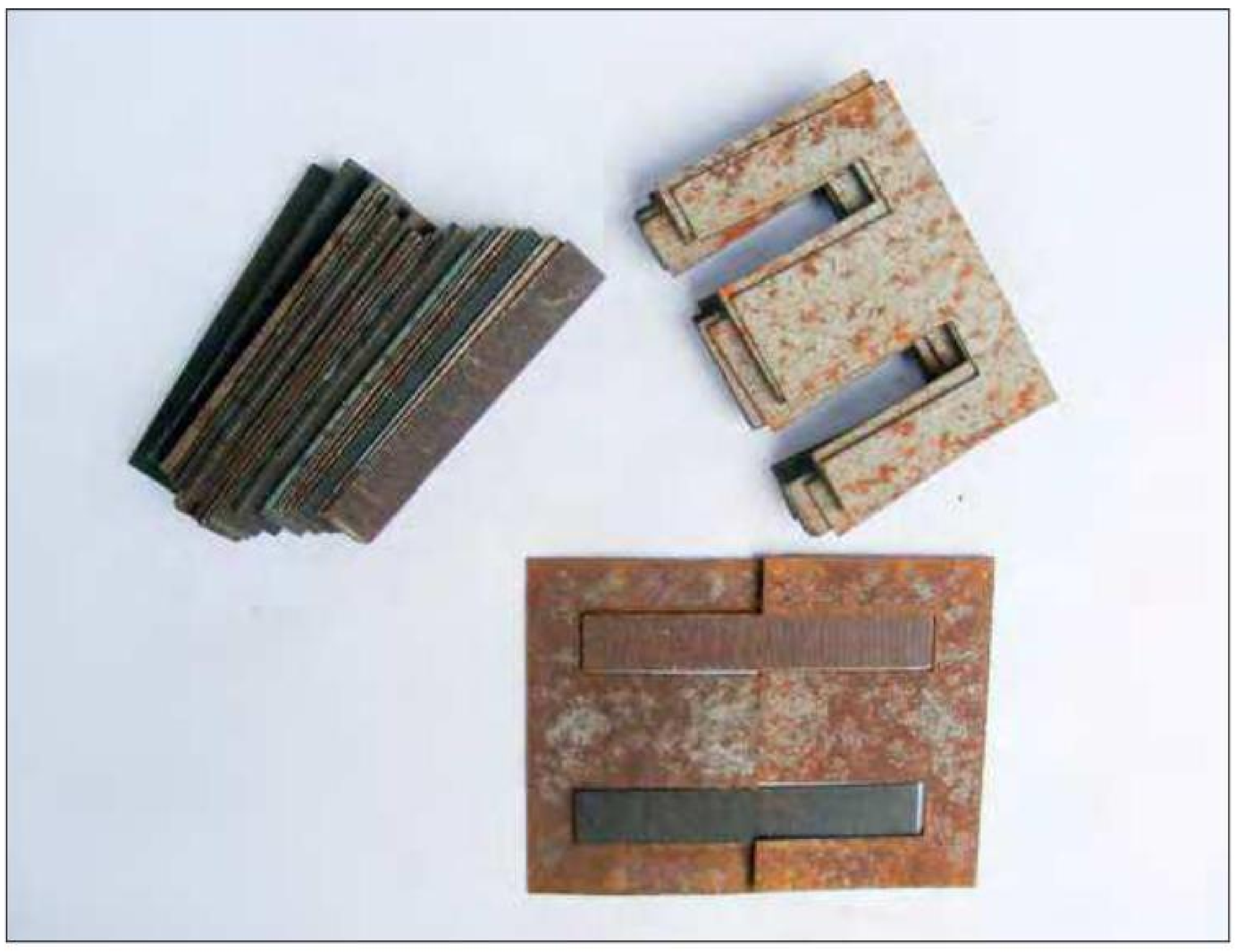


Fig.10. No-waste E and I shaped laminations. These are normally inserted alternately (interleaved) into the wound bobbin.

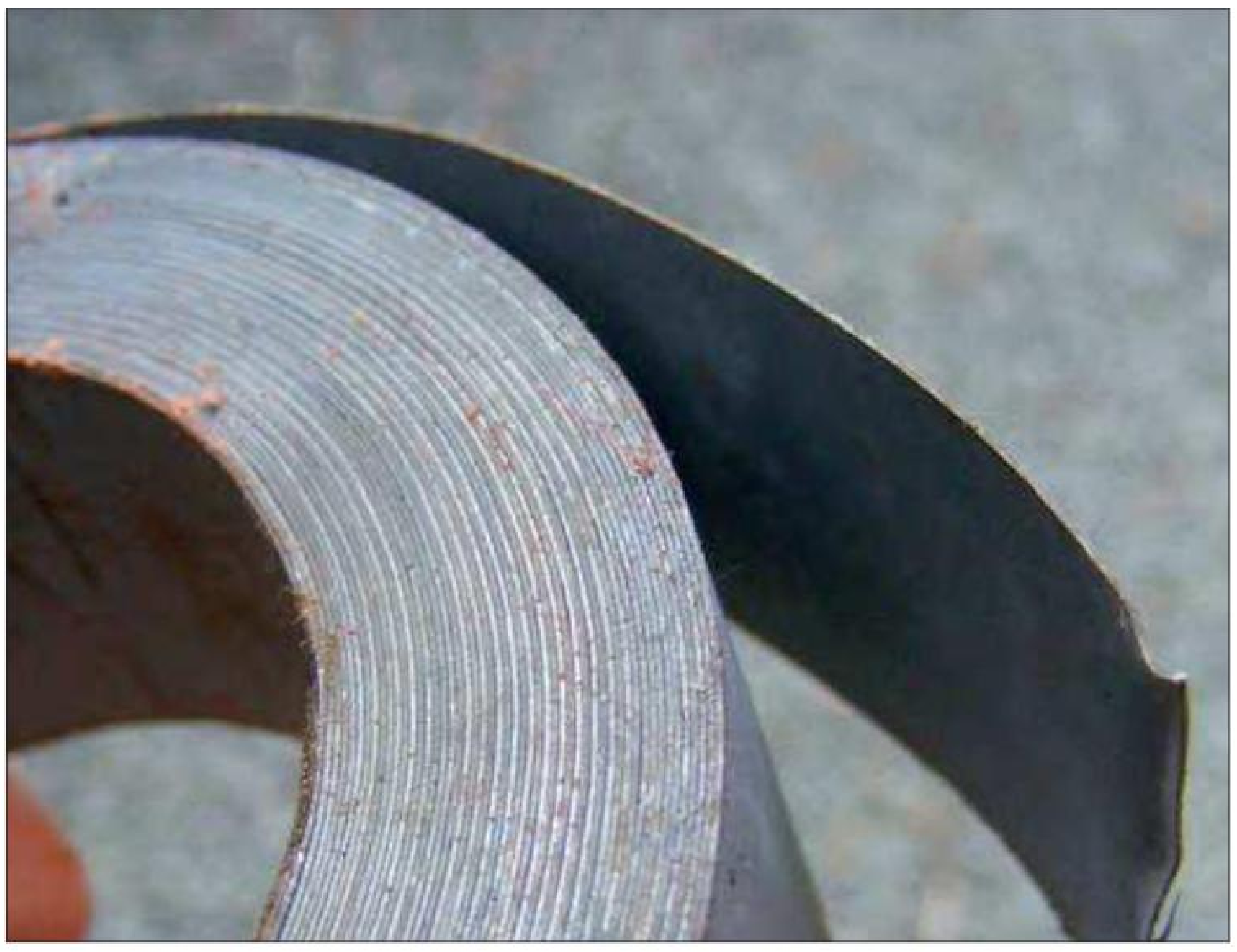


Fig.11. Toroidal transformer cores are wound from a long strip of steel into a ring.

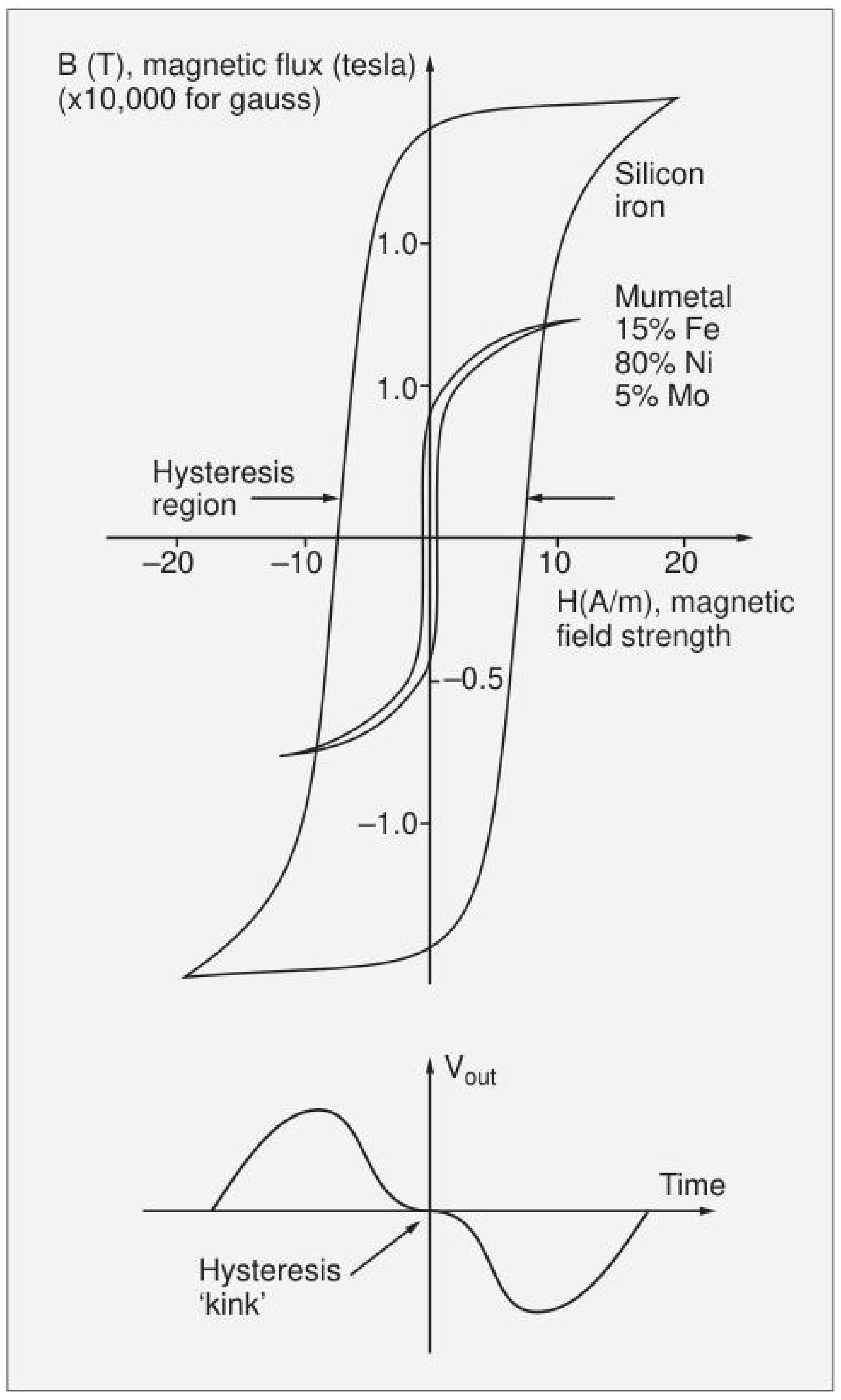


Fig.12. Magnetisation curve of silicon steel showing hysteresis kink. In power transformers this gives rise to a small heating loss. In signal transformers this gives a 'soft-sounding' third-harmonic distortion on low-level signals.

Eddy currents

Eddy currents arise because iron, the magnetic core of the transformer, is an electrical conductor and circulating currents are induced in it by the coils, causing heating. These can be mostly eliminated by laminating the steel into strips with an insulating coating. The flat strips allow the passage of magnetic flux but break up any eddy currents circulating in the conducting iron by reducing the cross-sectional area. The higher the frequency the thinner the laminations need to be. Westinghouse commercialised the production of transformers by developing a 'no-waste' cutting process of making E and I shaped laminations, as shown in Fig.10. These are interleaved into a ready-wound bobbin. Toroidal transformers use a strip of iron wound into a ring – see Fig.11. (It's only relatively recently that efficient machines have been developed to wind the wires efficiently through the core.) The steel is normally grain-orientated, where the crystals are all lined up the same way by the rolling process, which gives greater permeability in the direction of the flux.

Saturation

All *ferromagnetic* materials have a maximum level of magnetic flux density they can accommodate, above which little to no additional flux can be carried. At this point the material is said to be 'saturated', or 'in saturation'. It can be considered as a magnetic circuit form of clipping. For silicon steel (which is really magnetically soft iron, with around 4.5% silicon) this occurs at

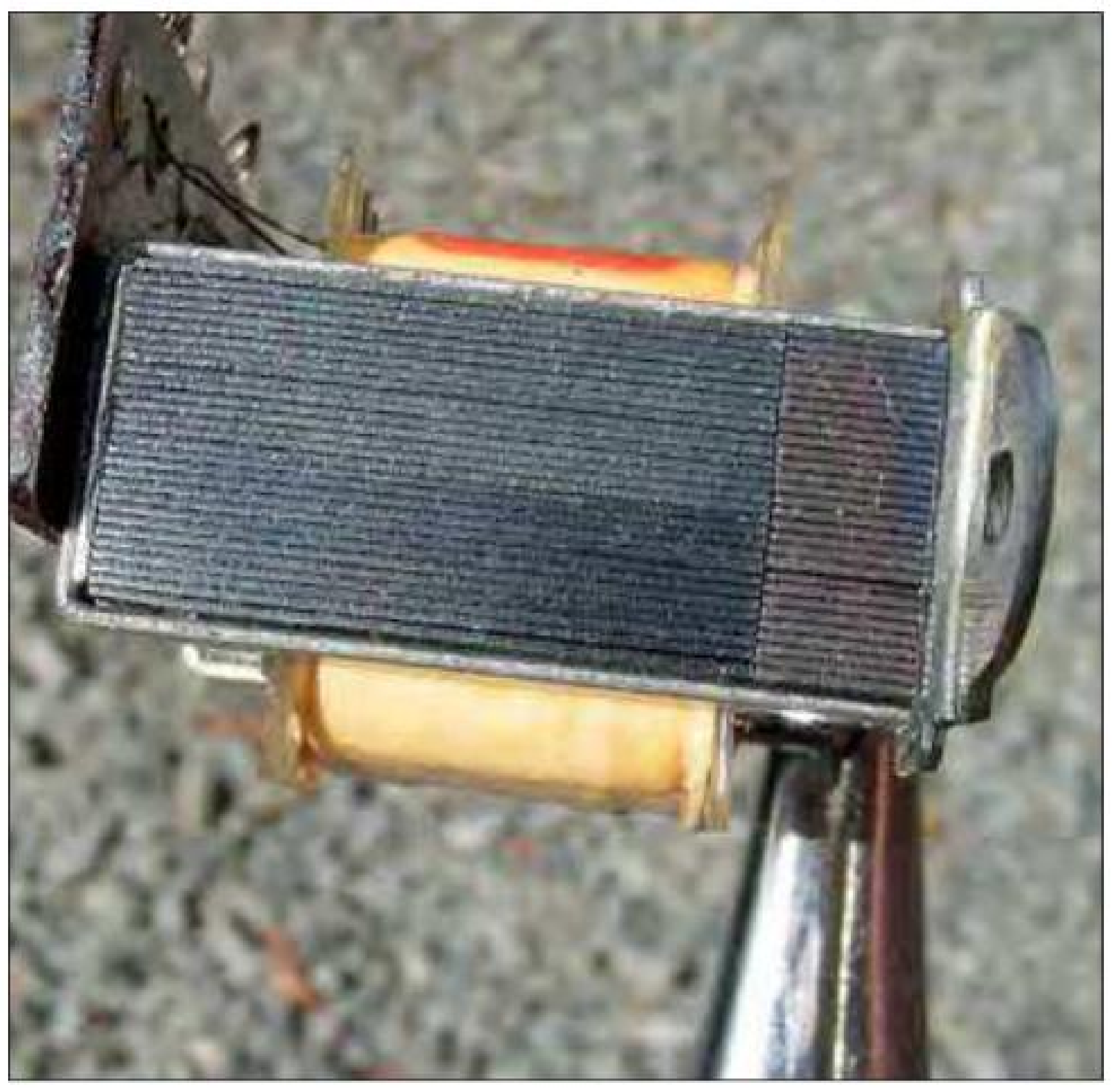


Fig.13. Butt jointing where all the Is are on one side rather than interleaving the laminations allows the transformer to accept much more DC current through the winding without saturation. This construction is employed for single-ended Class-A amplifiers and smoothing chokes. Toroidal transformers cannot accept any DC since they have no gap at all.

around 19,000 gauss. Mumetal, which is used for magnetic shielding saturates at only 8,500 gauss, and Radiometal at 16,000 gauss. When saturation occurs, the input impedance can suddenly drop, overloading the stage driving it. High-nickel-content materials such a Mumetal (80%) and Radiometal (50%) have higher permeability (ease of magnetisation) than silicon steel, so they

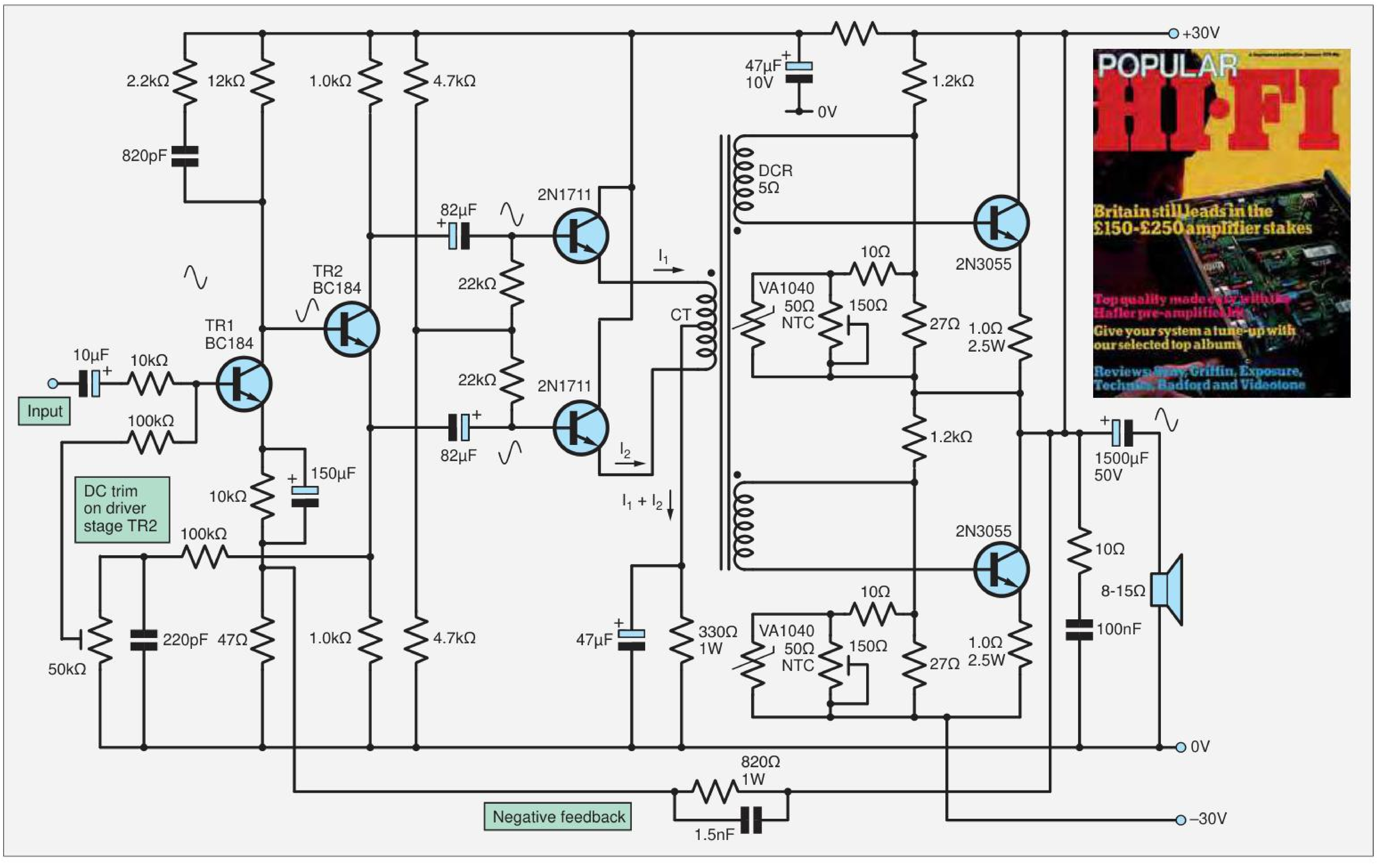

+30V
2.2kΩ
12kΩ
1.0kΩ
4.7kΩ
47µF
10V
0V
1.2kΩ
820pF
DCR
5Ω
82µF
2N1711
2N3055
I₁
10Ω
TR2
BC184
22kΩ
VA1040
50Ω
NTC
150Ω
CT
27Ω
1.0Ω
2.5W
TR1
BC184
10µF
10kΩ
22kΩ
2N1711
Input
100kΩ
82µF
I₂
1.2kΩ
1500µF
50V
DC trim
on driver
stage TR2
10kΩ
150µF
I₁ + I₂
100kΩ
2N3055
10Ω
10Ω
8-15Ω
50kΩ
220pF
47Ω
1.0kΩ
4.7kΩ
47µF
330Ω
1W
VA1040
50Ω
NTC
150Ω
27Ω
1.0Ω
2.5W
100nF
0V
820Ω
1W
Negative feedback
1.5nF
–30V
POPULAR
HI-FI
Britain still leads in the £150-£250 amplifier stakes
Give your system a tune-up with our selected top albums

Fig.14a. The circuit of the Rogers Ravensbourne amplifier used two emitter followers with their currents flowing in opposite directions to cancel DC magnetisation currents. The sound of this simple amplifier circuit was highly regarded in 1979. See inset photo Fig.14b.

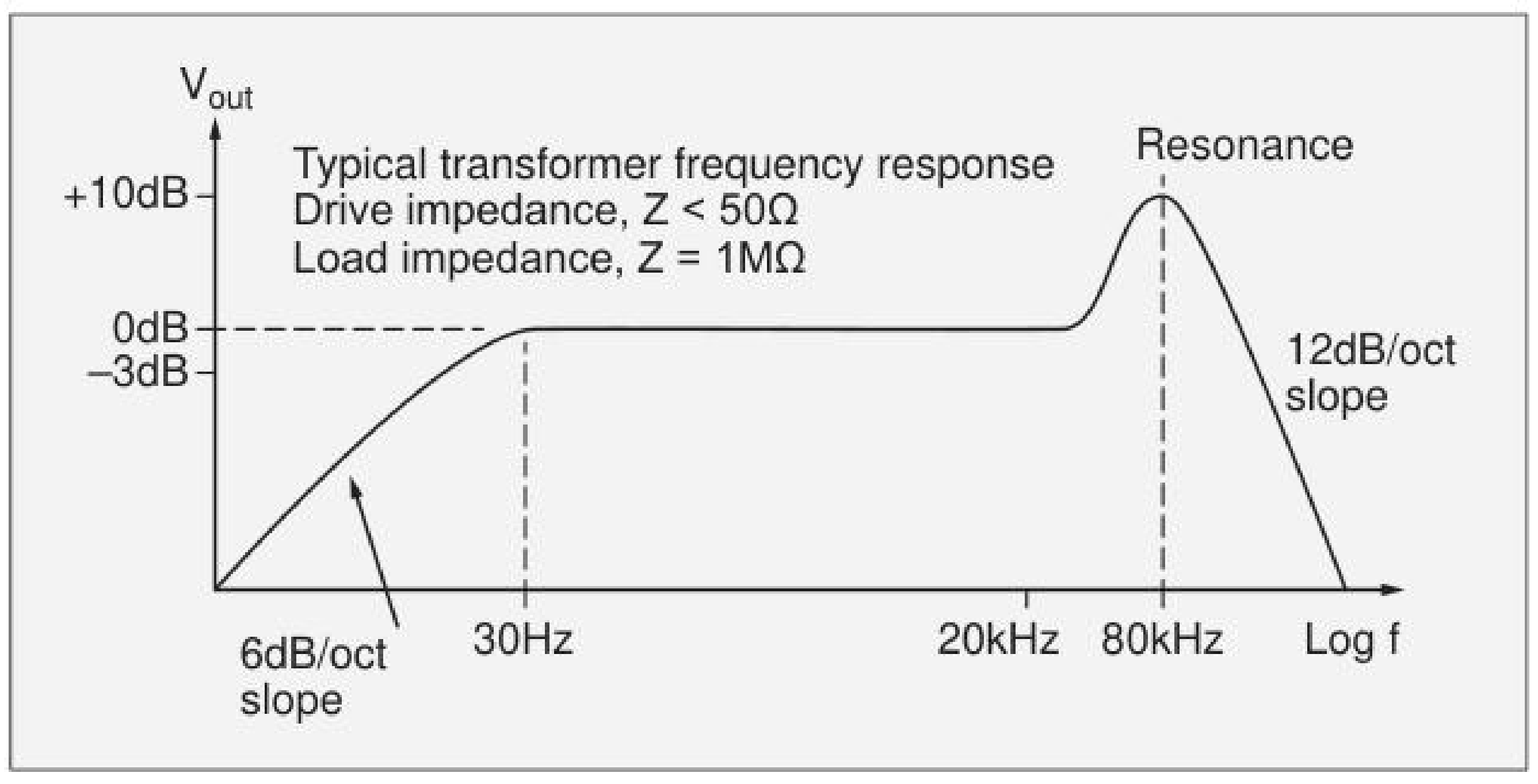


Fig.15. Typical frequency response of an audio transformer. A first-order bass roll-off below 30Hz and a hump at the high frequency second-order roll-off point.

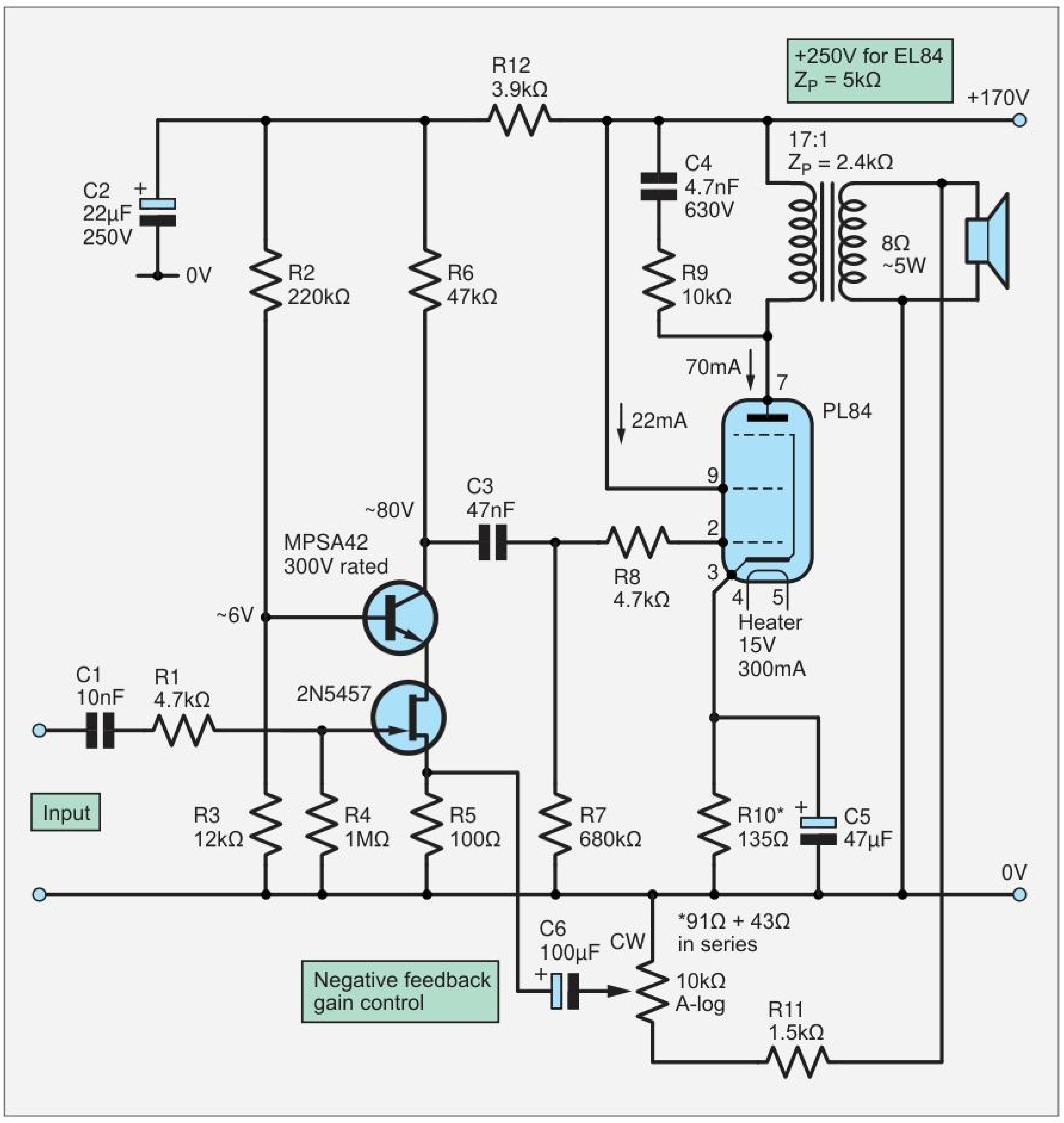

R12
3.9kΩ
+250V for EL84
Z_P = 5kΩ
+170V
C2
22µF
250V
0V
C4
4.7nF
630V
17:1
Z_P = 2.4kΩ
8Ω
~5W
R2
220kΩ
R6
47kΩ
R9
10kΩ
70mA
7
PL84
22mA
9
~80V
C3
47nF
2
MPSA42
300V rated
3
R8
4.7kΩ
4
5
Heater
15V
300mA
~6V
C1
10nF
R1
4.7kΩ
2N5457
Input
R3
12kΩ
R4
1MΩ
R5
100Ω
R7
680kΩ
R10*
135Ω
C5
47µF
0V
C6
100µF
CW
*91Ω + 43Ω
in series
Negative feedback gain control
10kΩ
A-log
R11
1.5kΩ

Fig.17. Small guitar practice amplifier using a PL84 valve. The PL84 is the cheap TV low-impedance version of the EL84 with a different 15V 300mA heater rather than the standard 6.3V.

need fewer turns for a given inductance, but the maximum flux density is less.

The maximum power and lowest frequency, ie the total magnetic flux of the system, that can be handled is determined by the physical size of the core. This is why big heavy transformers are needed for high powers. For audio output transformers, one old rule of thumb is 0.17lb (77g) of steel per watt.

Hysteresis

There is a kink in the BH curve of all magnetic materials due to hysteresis. This is where a reverse force is required to get the curve back to zero, as shown in Fig.12. This non-linearity causes the current drawn from the source to be distorted. Silicon steel has a wider hysteresis area than nickel-based alloys, so it produces higher distortion. Distortion can be minimised by driving the transformer with a low source impedance. It can be made lower still by using a negative impedance to cancel out the primary's series resistance. The lamination material can be identified by its appearance. Silicon steel has a dark grey finish and nickel-based laminations look bright and silvery.

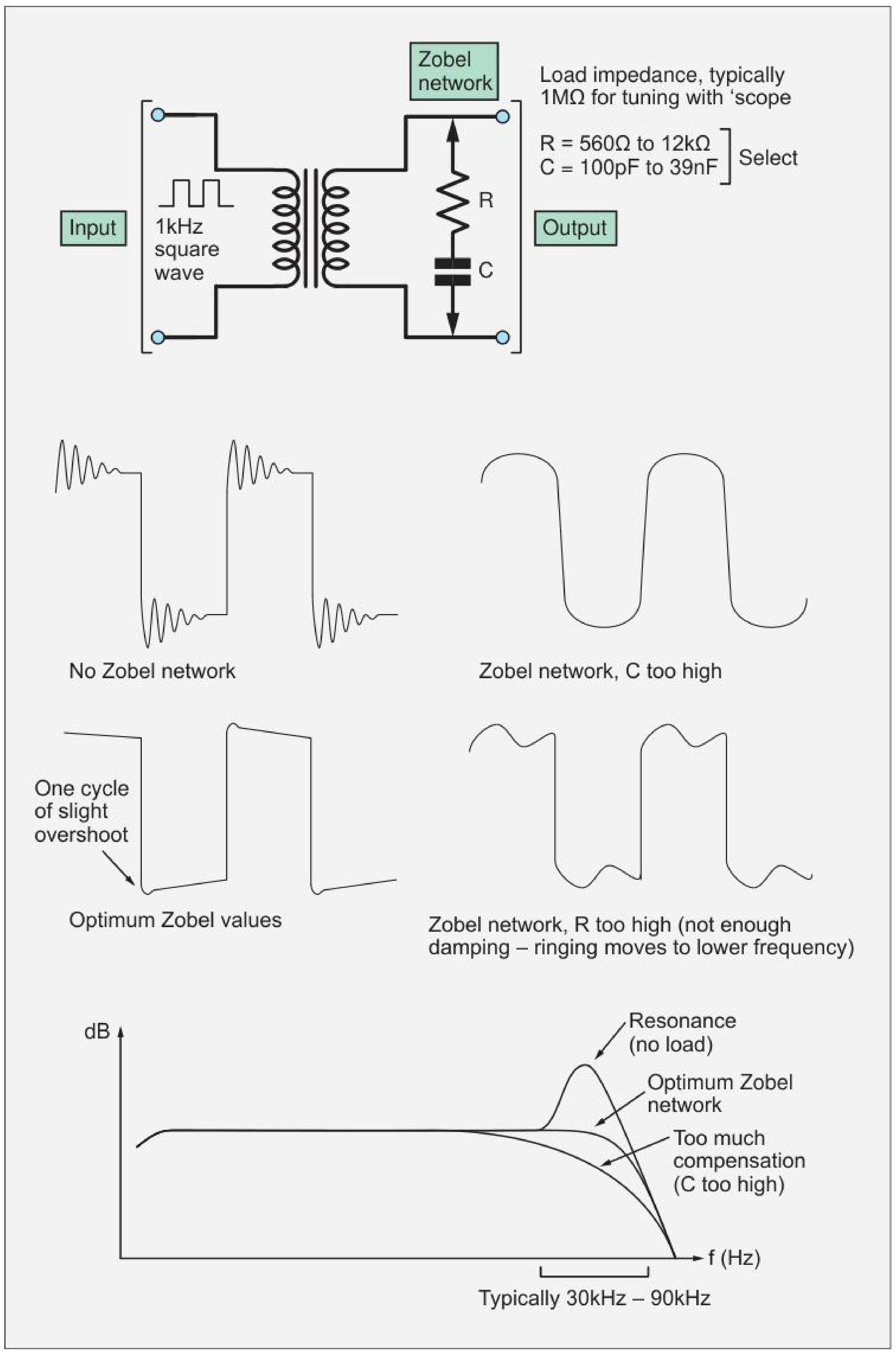

Zobel network
Load impedance, typically 1MΩ for tuning with 'scope
R = 560Ω to 12kΩ
C = 100pF to 39nF
Select
Input
1kHz square wave
R
C
Output
No Zobel network
Zobel network, C too high
One cycle of slight overshoot
Optimum Zobel values
Zobel network, R too high (not enough damping – ringing moves to lower frequency)
dB
Resonance (no load)
Optimum Zobel network
Too much compensation (C too high)
f (Hz)
Typically 30kHz – 90kHz

Fig.16. Adding a Zobel network damps the high frequency resonance. The values have to be determined empirically by feeding a 1kHz square wave into the transformer primary and tuning out the ringing on a 'scope.

DC current

Transformers do not like DC current in the windings as it eats up the device's saturation headroom. Magnetic induction voltage is proportional to rate of change of magnetic flux, so the lower the frequency the less efficient it gets. DC is zero frequency. If DC current has to be accommodated, such as in the output transformer of a single-ended class A amplifier, a 'resistance' needs to be inserted in the magnetic circuit to limit the flux which is analogous to electrical current. This is achieved by introducing a gap in the circuit. In the case of EI laminations, this can be achieved by a butt joint between the lumped 'E' and 'I' laminations, as shown in Fig.13. A plastic spacer can be inserted to increase the size of the gap if necessary. A side effect of this is to reduce the inductance. Clever circuits have been devised to cancel out the magnetisation of DC currents by running equal currents in opposite directions in the windings. This is one of the great benefits of push-pull operation. It was added to the driver transformers of the Bowes transistor amplifier and the Rogers Ravensbourne, as shown in Fig.14.

Capacitance and leakage inductance

Capacitive parasitics are responsible for causing high-frequency roll-off. At the upper end of the response curve these parameters often form a resonance (see Fig.15) similar to an

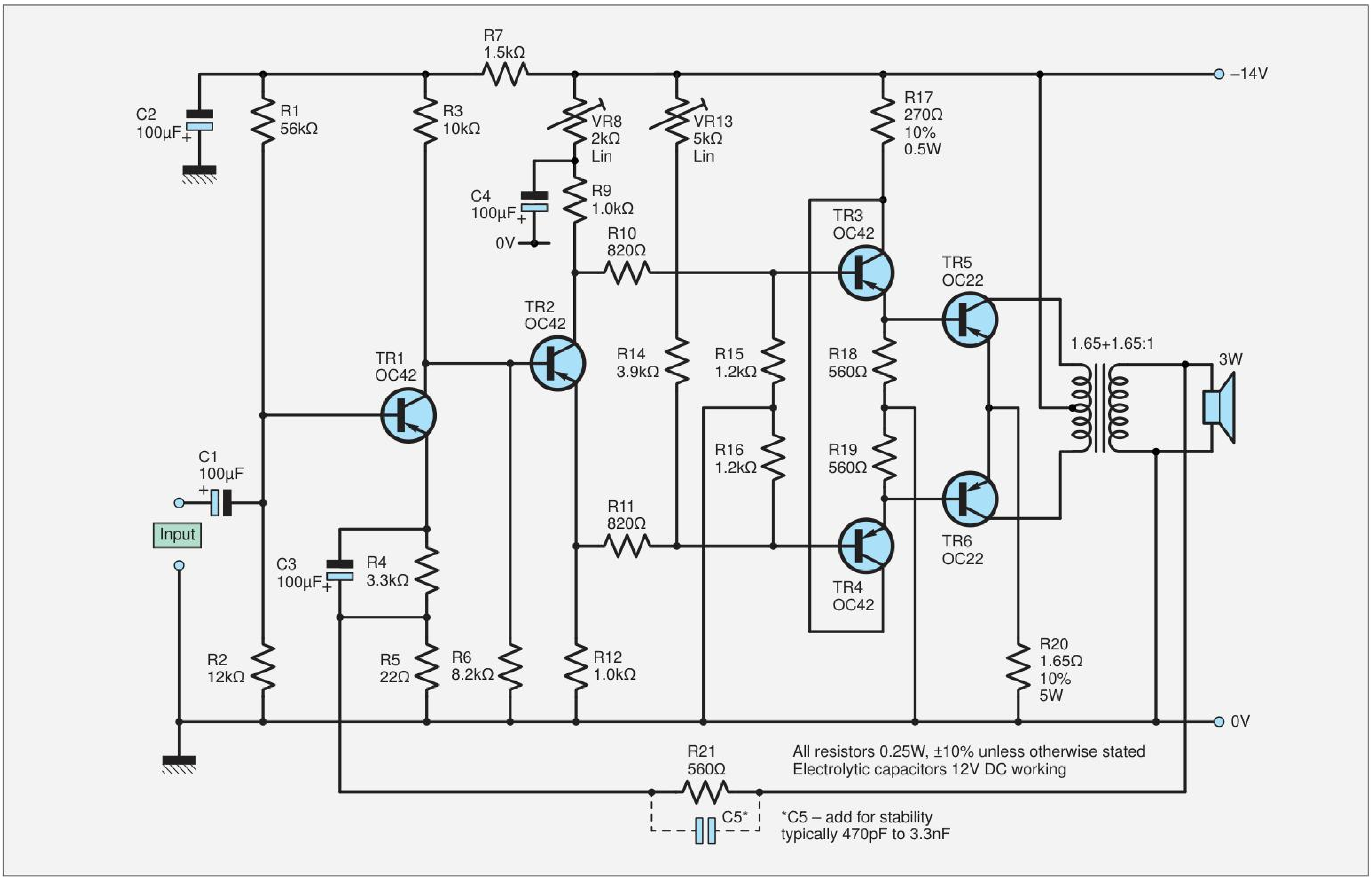
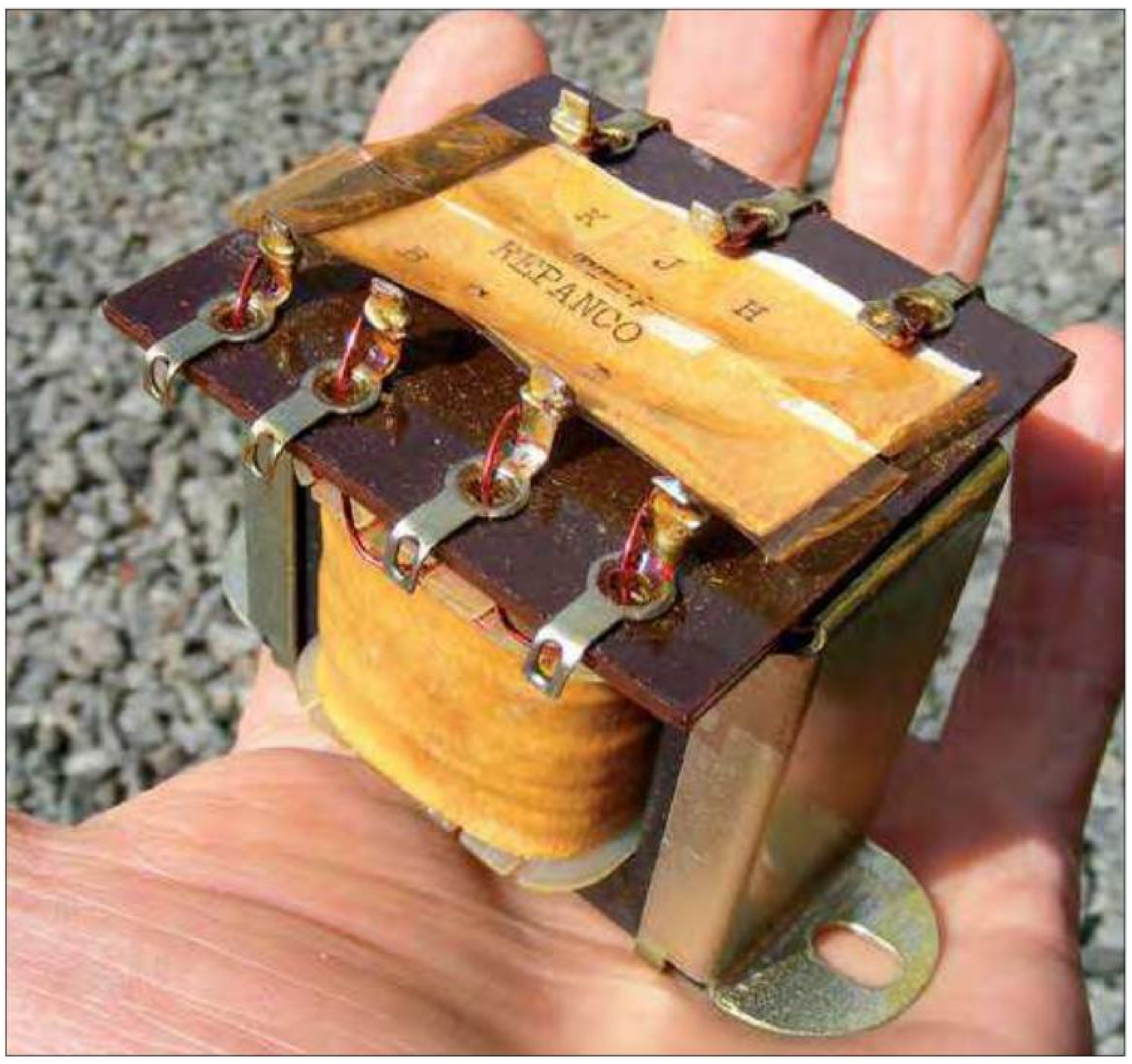

R7
1.5kΩ
–14V
C2
100µF
R1
56kΩ
R3
10kΩ
VR8
2kΩ
Lin
VR13
5kΩ
Lin
R17
270Ω
10%
0.5W
C4
100µF
R9
1.0kΩ
0V
R10
820Ω
TR3
OC42
TR5
OC22
TR2
OC42
TR1
OC42
R14
3.9kΩ
R15
1.2kΩ
R18
560Ω
1.65+1.65:1
3W
R16
1.2kΩ
R19
560Ω
C1
100µF
Input
R11
820Ω
TR6
OC22
C3
100µF
R4
3.3kΩ
TR4
OC42
R2
12kΩ
R5
22Ω
R6
8.2kΩ
R12
1.0kΩ
R20
1.65Ω
10%
5W
0V
R21
560Ω
All resistors 0.25W, ±10% unless otherwise stated
Electrolytic capacitors 12V DC working
C5*
*C5 – add for stability
typically 470pF to 3.3nF

Fig.18. a) (above) Mullard 5W class-A transistor amplifier circuit from 1961; b) (below) I have the output transformers and transistors for this design and will build one for its historical and teaching value. Apparently, they sounded quite good.


REPANCO
K
J
H

under-damped low-pass filter. This effect is usually suppressed with a series RC circuit called a Zobel network, named after the telecoms engineer who invented it – see Fig.16.

Output transformers

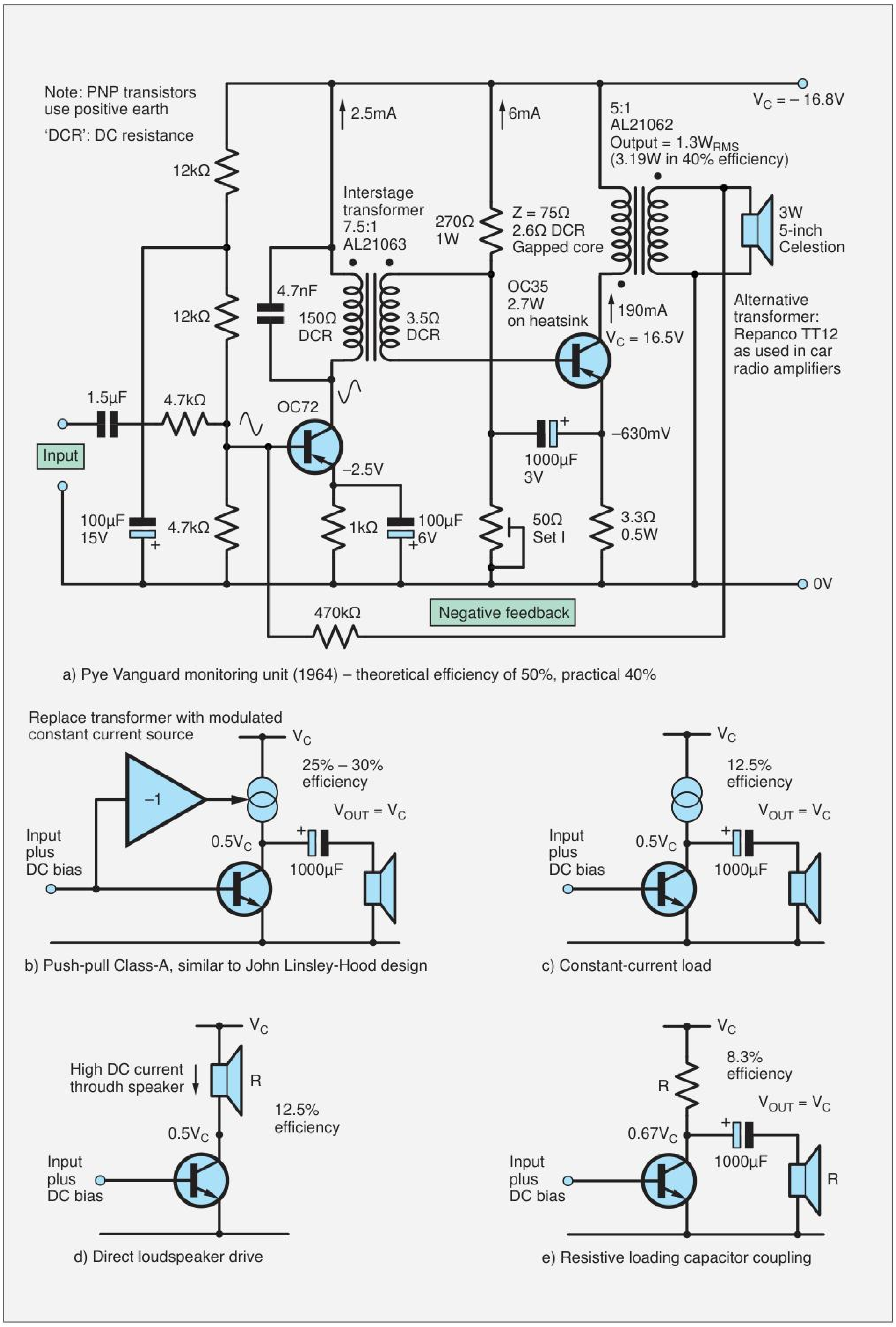
In pre-semiconductor days, valves were the only audio amplifying devices available. They are high-voltage (90 to 800V) low-current (10 to 100mA) devices, which means a high output impedance (Z_{out}) of around 1500 to 10,000Ω. Moving-coil loudspeakers are the opposite, having a low-impedance, typically 3 to 16Ω input impedance (Z_{in}). It is not possible to wind a reliable 5000Ω speaker voice-coil, so the only solution to this dilemma is an output matching transformer which transforms the load imposed on its secondary to a suitable one for the driving device – we looked at this last time. A typical transformer 'matching circuit' is shown below in Fig.17, a good valve amplifier using the trusted (and still available) EL84 output valve, which needs a 5kΩ load. This calls for a turns ratio of 40:1 to match a 3Ω loudspeaker. The result is an impedance ratio of 1600:1. Note that with a 3Ω loudspeaker this is reflected back to the primary, that is transformed to 1600 × 3 = 4800Ω. OEP do a suitable audio transformer.

If an 8Ω speaker were to be used then an impedance ratio of 625:1 would be needed, giving a load of 5kΩ. Working backwards, the square root of the impedance ratio is taken to get the turns ratio, giving 25:1. I once found a box of cheap PL84 valves used for old TVs. These have a relatively lower impedance, so when designing an output circuit around them I dropped the HT voltage and transformer ratio to 17:1 to produce a 2.4kΩ impedance.

As an interesting aside, early transistor amplifiers briefly used output transformers (Fig.18) but as their output current capability increased it became possible to drive loudspeakers directly.

Output transformers do tend to be big because they have to transmit significant power. A 20W transformer may weigh a couple of pounds (1kg). Cores are normally grain-orientated silicon iron for high power handling. Negative feedback is often applied around output transformers to reduce the distortion and widen the frequency response. However, there is a limit to the amount of feedback that can be applied due to phase shifts associated with the transformers high- and low-frequency roll-offs.

For class-A stages, output transformers enable the greatest theoretical efficiency to be achieved of 50%. (Fig.19a) Without a transformer, using a push-pull constant-current load the maximum possible efficiency drops to 25%. With a fixed current source or a simple loudspeaker load it's 12.5% and


Note: PNP transistors use positive earth
'DCR': DC resistance
12kΩ
12kΩ
2.5mA
6mA
5:1
AL21062
Output = 1.3W$_{RMS}$
(3.19W in 40% efficiency)
V_C = – 16.8V
Interstage transformer 7.5:1 AL21063
270Ω 1W
Z = 75Ω
2.6Ω DCR
Gapped core
3W
5-inch
Celestion
4.7nF
150Ω DCR
3.5Ω DCR
OC35
2.7W
on heatsink
190mA
V_C = 16.5V
Alternative transformer: Repanco TT12 as used in car radio amplifiers
1.5µF
4.7kΩ
OC72
Input
–2.5V
1000µF 3V
–630mV
100µF 15V
4.7kΩ
1kΩ
100µF 6V
50Ω Set I
3.3Ω 0.5W
0V
470kΩ
Negative feedback
Replace transformer with modulated constant current source
V_C
–1
25% – 30% efficiency
V_{OUT} = V_C
Input plus DC bias
0.5V_C
1000µF
12.5% efficiency
High DC current throudh speaker
R
0.5V_C
8.3% efficiency
0.67V_C

a) Pye Vanguard monitoring unit (1964) – theoretical efficiency of 50%, practical 40%

b) Push-pull Class-A, similar to John Linsley-Hood design

c) Constant-current load

d) Direct loudspeaker drive

e) Resistive loading capacitor coupling

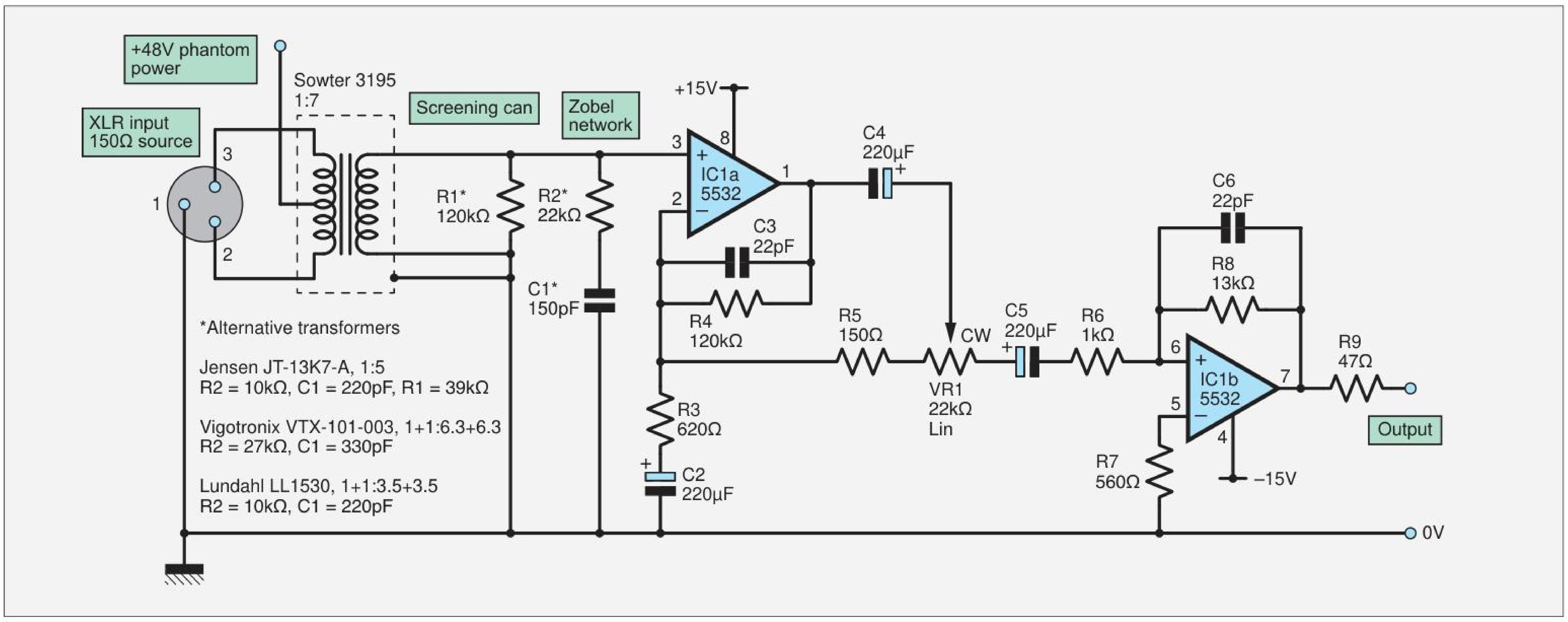
Fig.19. For single ended Class-A amplifiers, transformers provide the most efficient way of coupling the loudspeaker. a) Transformer coupled – 1960s car radio amplifier, theoretically 50% efficient. b) No transformer, modulated constant-current load. Efficiency 25-30% due to quarter of power rail voltage lost across current source, c) unmodulated constant-current load, 12.5% efficiency, d) direct loudspeaker connection, 12.5% efficiency since half the supply voltage lost. e) RC coupled, only 8.3% efficient because audio is also dissipated in the power resistor. All figures just before clipping. At lower volumes it's even worse.

with RC coupling (Fig.19e), it's a paltry 8.3% with half the audio power going into the RC network resistor. The high efficiency in the transformer circuit occurs because the transformer has a high AC load resistance and a small DC resistance. This means the output device sees the full power-rail voltage and the resulting AC swing is twice that.

Input transformer

Moving to the 'other end' of an audio circuit, a common use for audio transformers is impedance matching on the input. A common example is a preamplifier for moving-coil microphones (which are typically 600Ω source impedance). The classic NE5534 op amp likes to see a source impedance of 4.5kΩ for lowest noise, so a step-up transformer of around 1:7 is requird – see the circuits in Fig.20.

Input transformers only need to be capable of transmitting a couple of milliwatts before saturation, so they tend to be quite small and employ Mumetal cores. A microphone input transformer is shown in Fig.21. They typically have turns ratios in the region of 4:1 to 12:1. Input transformers are also used to


+48V phantom power
XLR input 150Ω source
Sowter 3195 1:7
Screening can
Zobel network
+15V
3
1
2
R1* 120kΩ
R2* 22kΩ
C1* 150pF
IC1a 5532
8
1
C3 22pF
R4 120kΩ
C4 220µF
R3 620Ω
C2 220µF
R5 150Ω
CW
VR1 22kΩ Lin
C5 220µF
R6 1kΩ
C6 22pF
R8 13kΩ
IC1b 5532
6
5
7
4
R7 560Ω
–15V
R9 47Ω
Output
0V

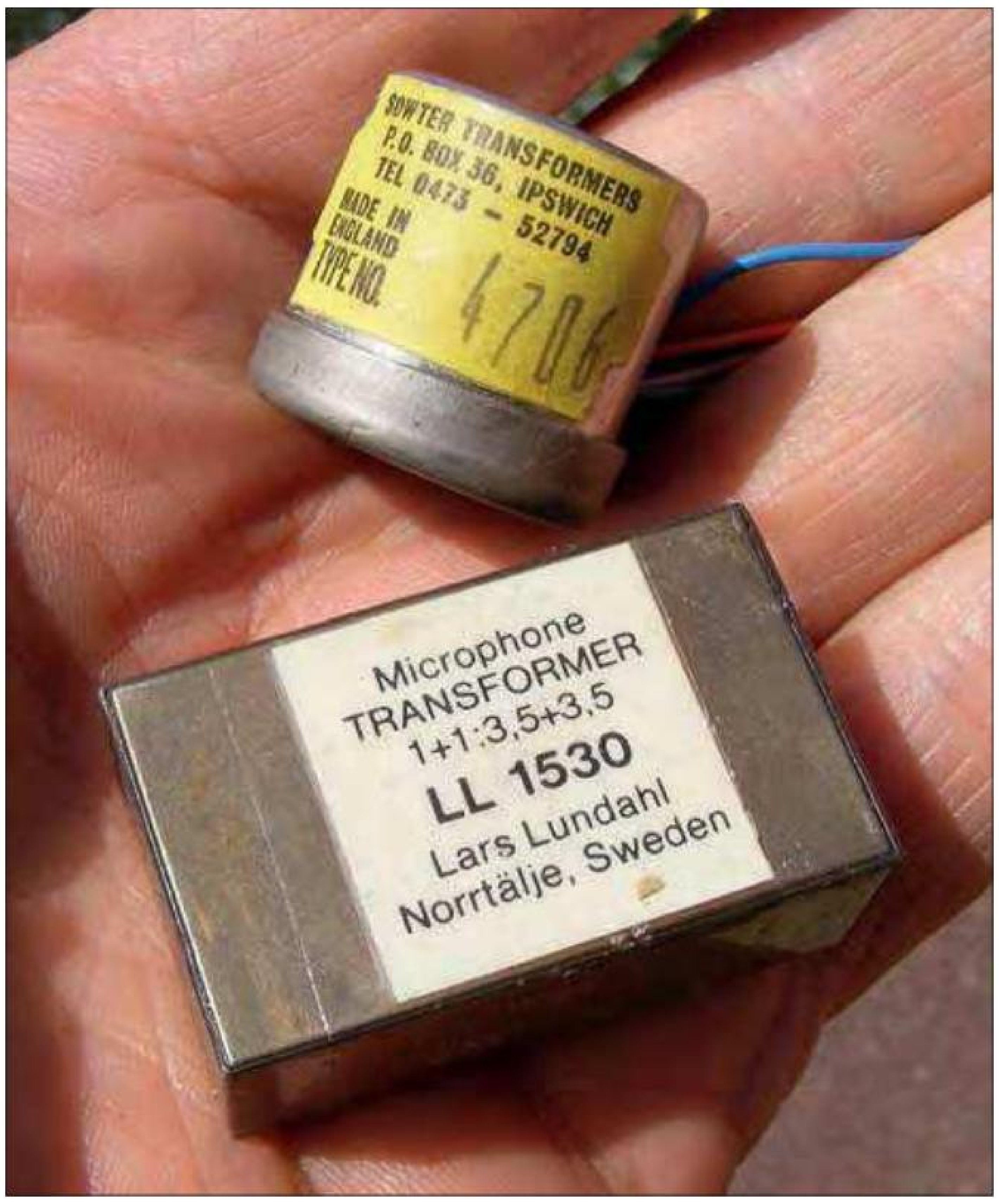
*Alternative transformers

Jensen JT-13K7-A, 1:5
R2 = 10kΩ, C1 = 220pF, R1 = 39kΩ

Vigotronix VTX-101-003, 1+1:6.3+6.3
R2 = 27kΩ, C1 = 330pF

Lundahl LL1530, 1+1:3.5+3.5
R2 = 10kΩ, C1 = 220pF

Fig.20. Transformer-coupled microphone pre-amp. Note how a centre-tap can provide a lossless phantom power connection. Whole circuit gain: +14dB to +70dB. (Based on a Steve Dove design).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P.O. BOX 36, IPSWICH
TEL 0473 – 52794
MADE IN ENGLAND
TYPE NO.
4706
Microphone
TRANSFORMER
1+1:3,5+3,5
LL 1530
Lars Lundahl
Norrtälje, Sweden

Fig.21. Typical microphone transformers. Note the round Mumetal can on the Sowter transformer. The Swedish Lundahl, along with American Jensen transformers are possibly the best.

create interference-rejecting balanced inputs. These low-level transformers are enclosed in Mumetal cans to shield them from hum.

Interstage transformers

When amplifying devices were very expensive it was essential to maximise the gain from each stage, so an impedance-matching 'interstage transformer' was employed between stages. These disappeared from valve amplifiers around 1940, only to reappear again in the early days of expensive transistors. These transformers were also called driver transformers, where dual secondaries were used to provide the phase splitting for push-pull output stages. A medium level of power is needed, around 20-200mW, so Radiometal is normally employed for lowest distortion. A common interstage transformer turns ratio range of between 1.5:1 to 5:1 was used for simple transistor pre-amplifiers such as the circuit shown in Fig.22. They are always step down to match the high-impedance collector output with the low-impedance base input. The low DC resistance of the transformer also provides a perfect path for any base leakage currents resulting in stable bias conditions. This circuit (shown built) in Fig.23 was used as an educational board – it makes a fantastic fuzz box!

Next month, we'll build something! – some transformer mounting PCBs.

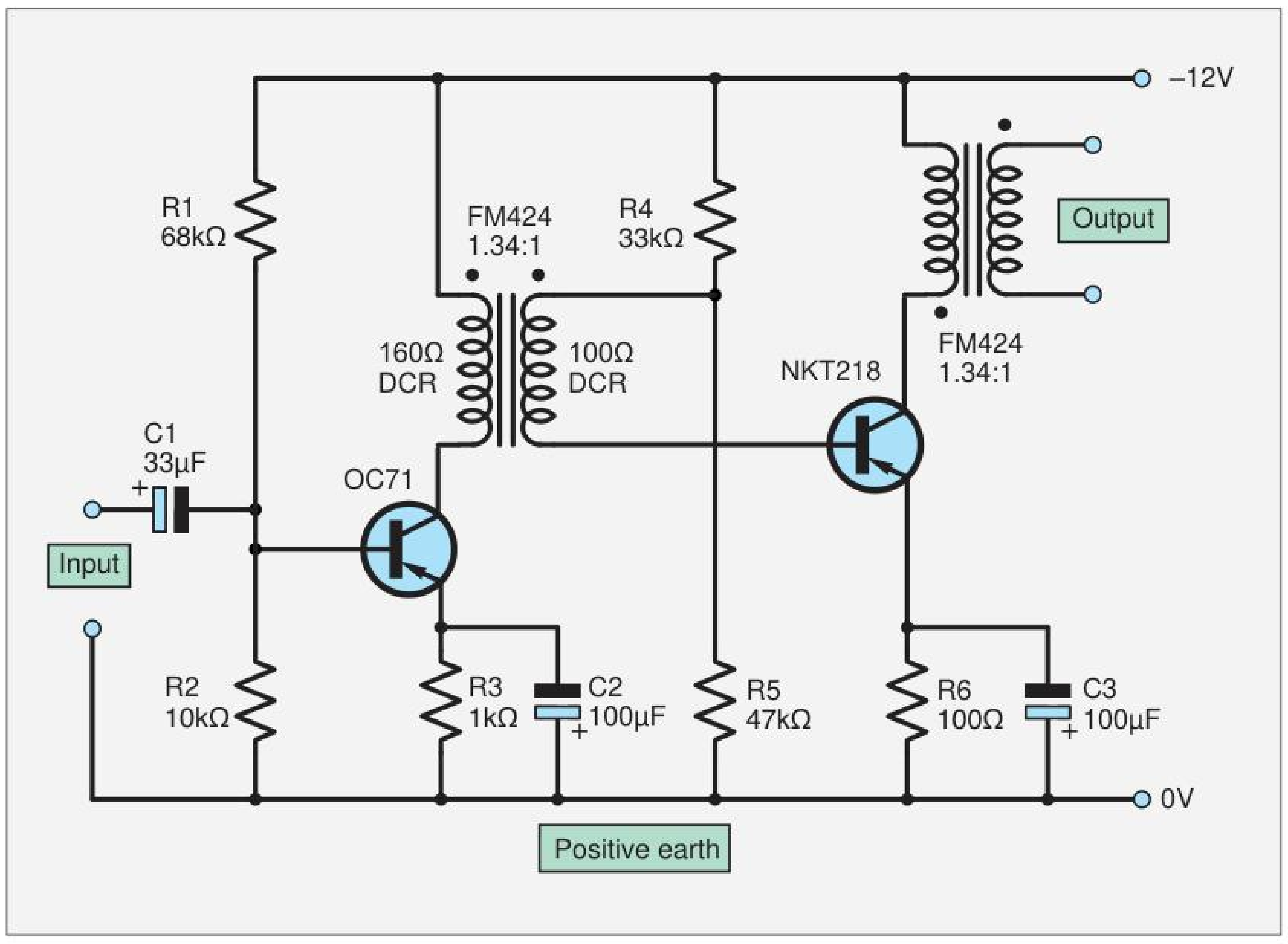
–12V
R1
68kΩ
FM424
1.34:1
R4
33kΩ
Output
FM424
1.34:1
160Ω
DCR
100Ω
DCR
NKT218
C1
33µF
OC71
Input
R2
10kΩ
R3
1kΩ
C2
100µF
R5
47kΩ
R6
100Ω
C3
100µF
0V
Positive earth

Fig.22. Lan-Elec Ltd audio-frequency No.5, two-stage transformer-coupled pre-amplifier.

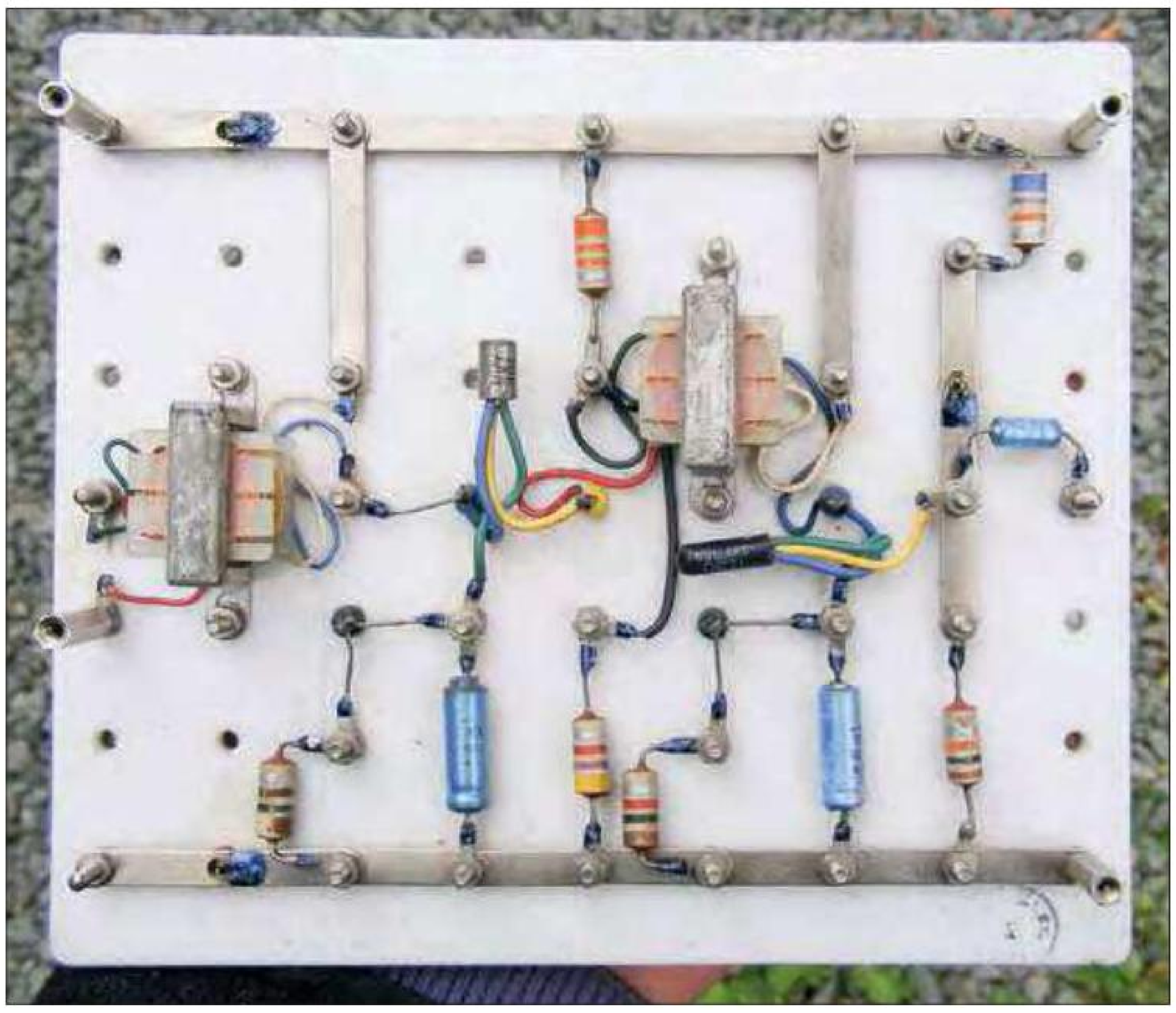

Fig.23. The circuit in Fig.22 was used as a teaching aid in an old Radio and TV servicing college in South London in the 1970s. Now redeployed as a germanium fuzz box.

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AUGUST 2022

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Simple MIDI Music Keyboard (for 8 switches)	23101213	6.95
Nano Pong	08105212	7.95

JULY 2022

PROJECT	CODE	PRICE
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JUNE 2022

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PIC Programming Helper for 8, 14 or 20-pin PICs	24106212	10.95
Advanced GPS Computer	05102211	9.95

MAY 2022

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Complete set of 14 PCBs for Analogue Vocoder	AO2-MAY22	97.95
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Programmed PIC for Digital FX Unit using potentiometer	FX2-MAY22	8.95

APRIL 2022

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Digital FX Unit – using potentiometer	01102211	9.95
Digital FX Unit – using BCD switch	01102212	9.95
Universal Audio PSU	AO1-APR22	11.95

MARCH 2022

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Busy Loo Indicator	16112201	£5.95
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FEBRUARY 2022

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JANUARY 2022

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MiniHeart: A Miniature Heartbeat Simulator	01109201	8.95

DECEMBER 2021

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NOVEMBER 2021

PROJECT	CODE	PRICE
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OCTOBER 2021

PROJECT	CODE	PRICE
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SEPTEMBER 2021

PROJECT	CODE	PRICE
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Audio DDS Oscillator rotary encoder	01110201-ENC	6.95
Programming Adaptor Board for Audio DDS Oscillator	01110202	£5.95
High-power Ultrasonic Cleaner main PCB	04105201	£14.95
High-power Ultrasonic Cleaner front-panel PCB	04105202	
Night Keeper Lighthouse PCB	08110201	£5.95

AUGUST 2021

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Ol' Timer 8x8 RGB LED module using WS2812B	19104201-88	£8.95
Ol' Timer set of acrylic case pieces and spacer	19104201-ACR	£8.75
Ol' Timer DS3231 RTC IC wide SOIC-16	19104201-RTC	£5.95
Wideband Digital RF Power Meter	04106201	£9.75
Switchmode 78xx regulators ***(PACK of 5!)***	18105201	£7.95
Cool Beans SMAD display	CB-AUG21	£11.95

JULY 2021

PROJECT	CODE	PRICE
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IR Remote Control Assistant (Jaycar version)	15005201	£8.95
IR Remote Control Assistant (Altronics version)	15005202	£8.95
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Touchscreen Wide-range RCL Box (Ind/Cap module)	04104202	
KickStart Part 3 – Gyrator-based Audio Filter	KS3-2021	£7.95

MAY 2021

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APRIL 2021

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MARCH 2021

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FEBRUARY 2021

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JANUARY 2021

PROJECT	CODE	PRICE
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DECEMBER 2020

PROJECT	CODE	PRICE
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Always check price and availability in the latest issue or online. A large number of older boards are listed for ordering on our website.

In most cases we do not supply kits or components for our projects. For older projects it is important to check the availability of all components before purchasing PCBs.

Back issues of articles are available – see Back Issues page for details.

Double-sided | plated-through holes | solder mask

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45V/8A Power Supply front panel five-way display bezel	18111181-BZ	£3.95
Five-way LCD Panel Meter/Display	18111182	£7.95
OCTOBER 2020		
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Precision Signal Amplifier	04107191	£7.95
SEPTEMBER 2020		
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PE Theremin PSU transformer	AO-0920-02	£7.95
Micromite Explore-28	07108191	£6.95
Ultrabrite LED Driver	16109191	£6.95
AUGUST 2020		
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PE Mini-organ selected parts	AO-0720-02	£8.95
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Teach-In 2017

Introducing the BBC micro:bit

PYTHON CODING ON THE BBC MICRO:BIT

Jim Gatenby

Python is the leading programming language, easy to learn and widely used by professional programmers. This book uses MicroPython, a version of Python adapted for the BBC Micro:bit.

Among the many topics covered are: main features of the BBC micro:bit including a simulation in a web browser screen; various levels of programming languages; Mu Editor for writing, saving and retrieving programs, with sample programs and practice exercises; REPL, an interactive program for quickly testing lines of code; scrolling messages, creating and animating images on the micro:bit's LEDs; playing and creating music, sounds and synthesized speech; using the on-board accelerometer to detect movement of the micro:bit on three axes; glossary of computing terms.

This book is written using plain English, avoids technical jargon wherever possible and covers many of the coding instructions and methods which are common to most programming languages. It should be helpful to beginners of any age, whether planning a career in computing or writing code as an enjoyable hobby.

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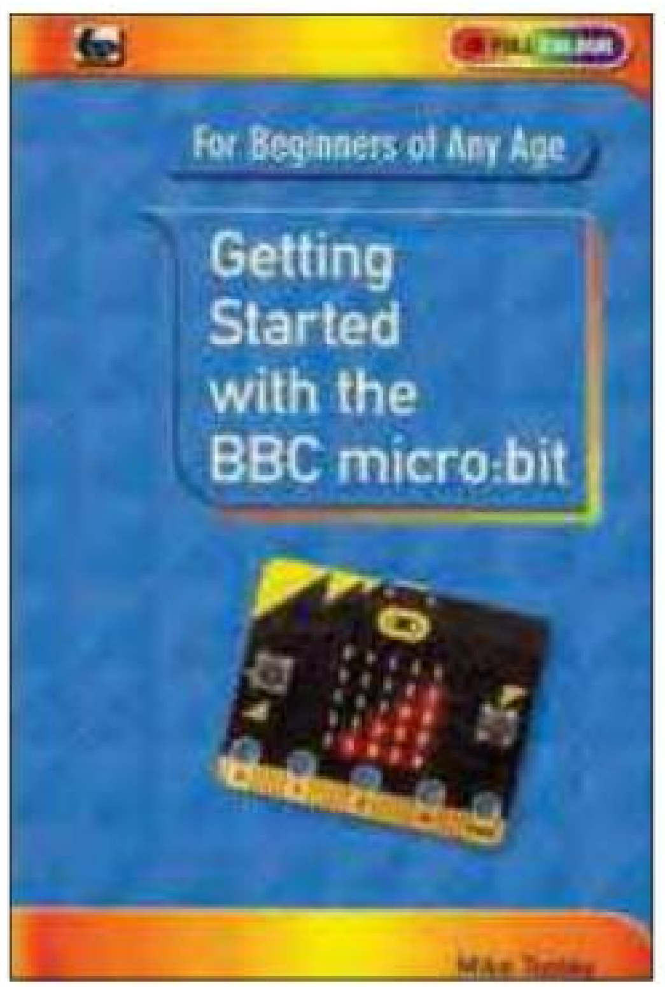

GETTING STARTED WITH THE BBC MICRO:BIT

Mike Tooley

Not just an educational resource for teaching youngsters coding, the BBC micro:bit is a tiny low cost, low-profile ARM-based single-board computer. The board measures 43mm × 52mm but despite its diminutive footprint it has all the features of a fully fledged microcontroller together with a simple LED matrix display, two buttons, an accelerometer and a magnetometer.

Mike Tooley's book will show you how the micro:bit can be used in a wide range of applications from simple domestic gadgets to more complex control systems such as those used for lighting, central heating and security applications. Using Microsoft Code Blocks, the book provides a progressive introduction to coding as well as interfacing with sensors and transducers.

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John Nussey

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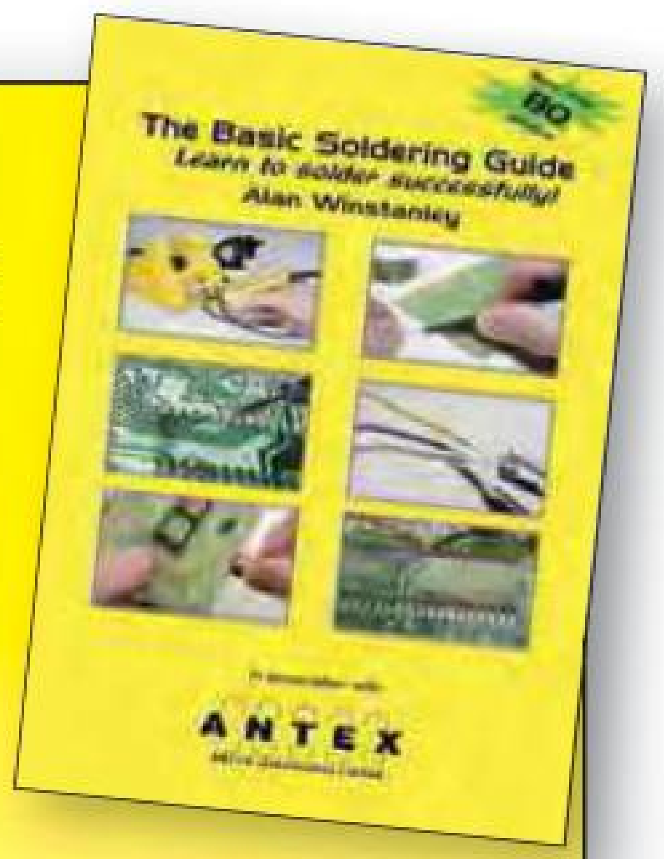

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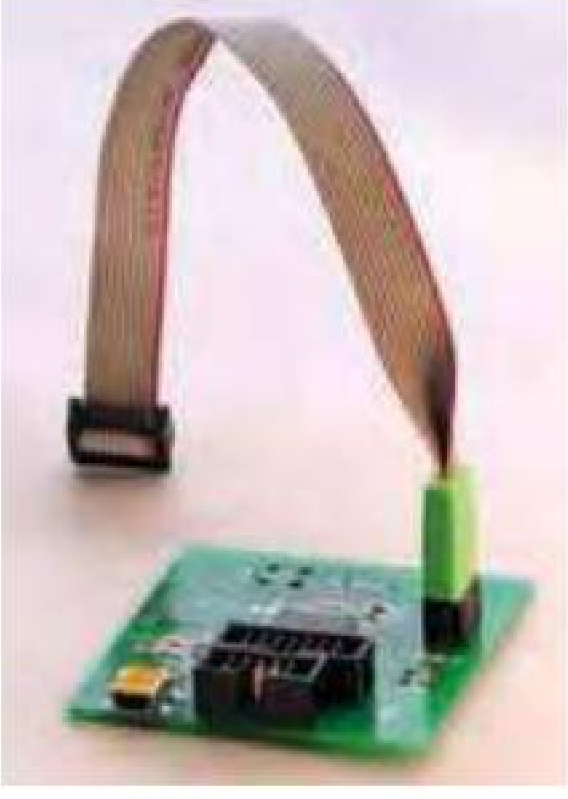

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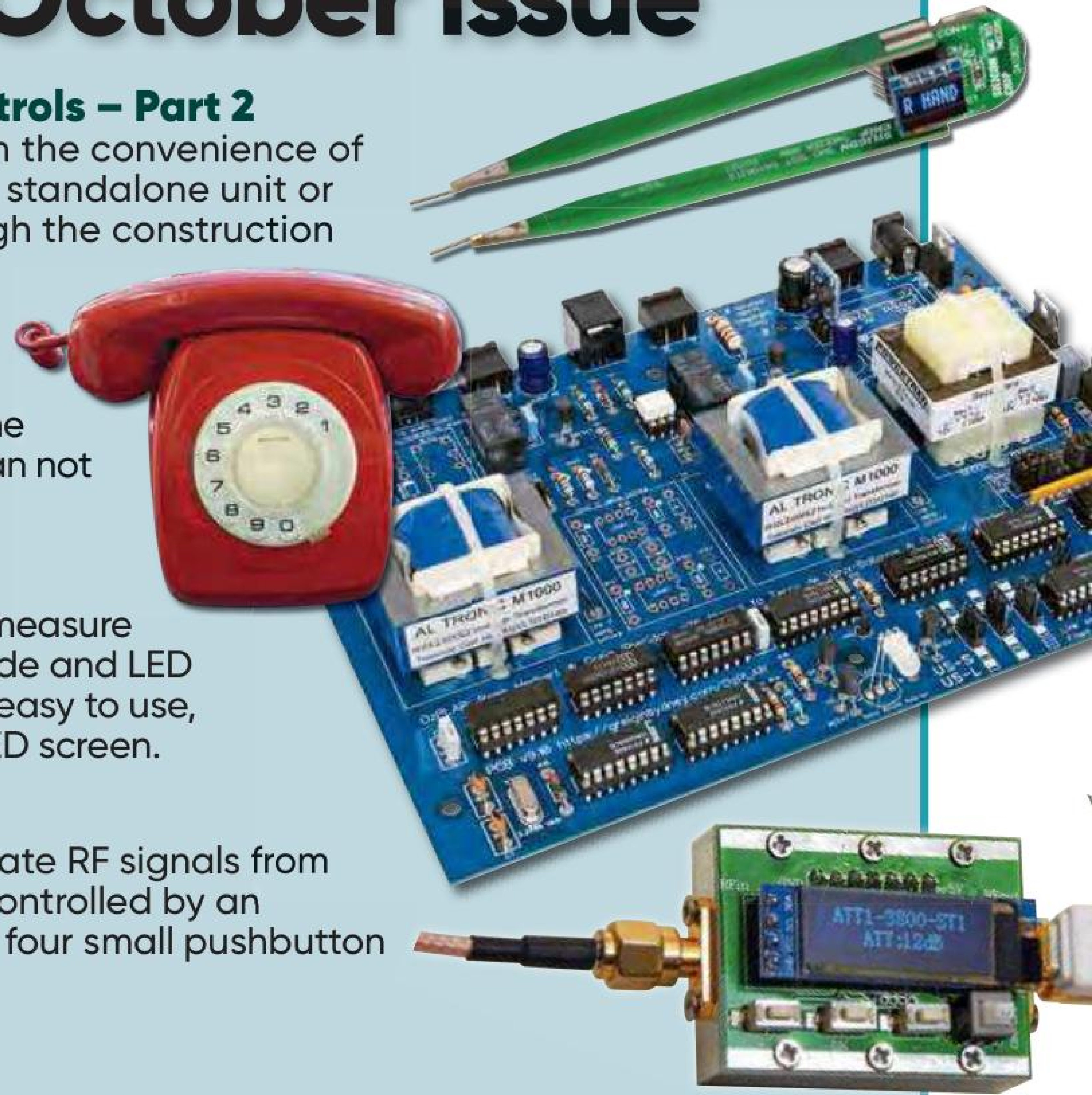

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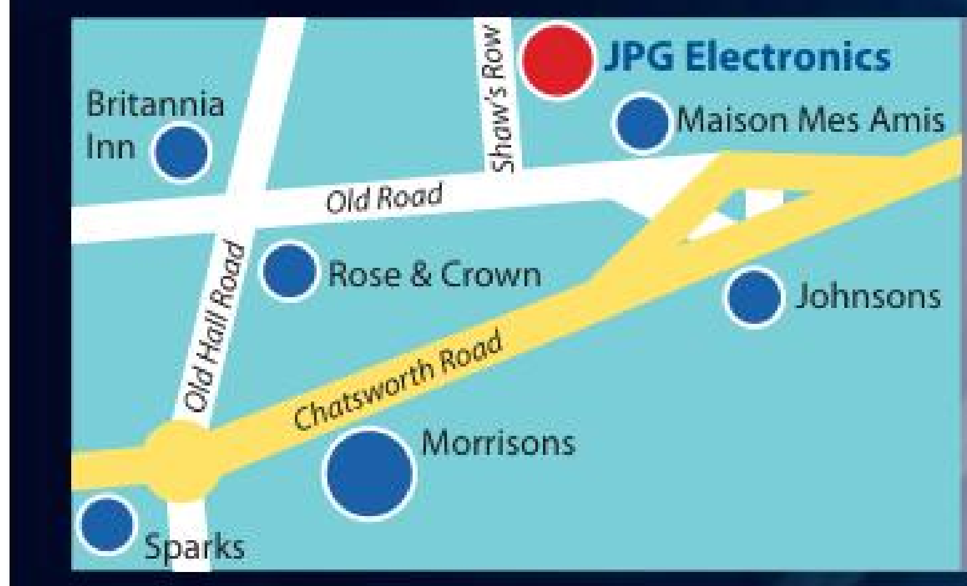

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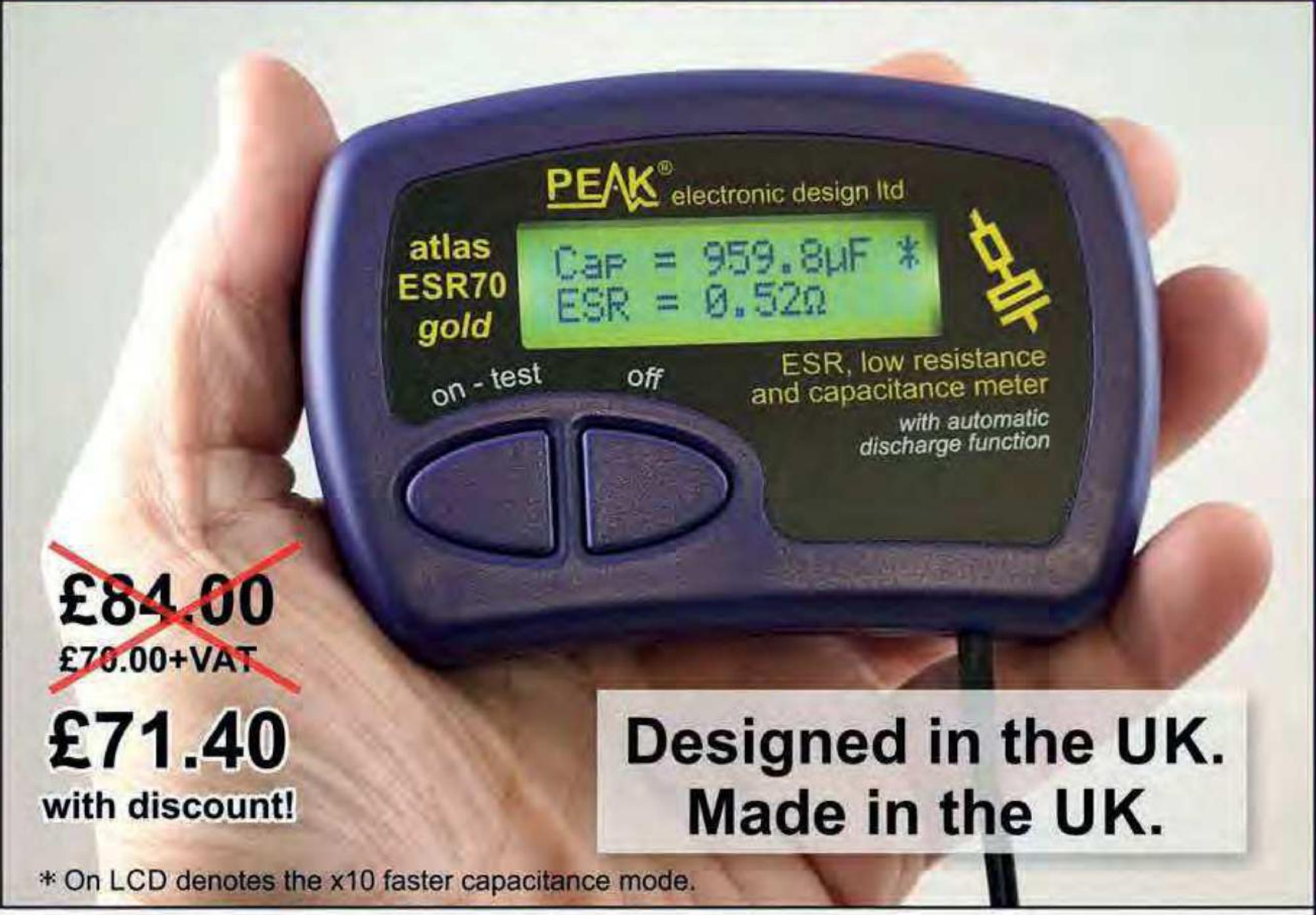

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* On LCD denotes the x10 faster capacitance mode.

What does the new ESR70 Gold do?

The new improved **ESR70 Gold** is great for identifying capacitors (especially electrolytics) that may be suffering from high ESR.

ESR is a great indicator of capacitor condition, especially in applications demanding high ripple currents such as reservoir capacitors, SMPSUs, motor capacitors etc.

The **ESR70 Gold** will even perform a controlled discharge of your caps!

ESR Analysis starts automatically when you apply the probes to the capacitor, even in-circuit.

For out-of-circuit testing, the **ESR70 Gold** also measures capacitance more rapidly than ever.

This **ESR70 Gold** includes removable gold-plated croc-clip probes, long ultra-flexible silicone test leads, alkaline AAA battery and a very comprehensive user guide.

All Peak products carry a 24 month warranty. Even after that period, we offer very low cost repair, update and calibration services.

Gold Features

- Much wider capacitance range 0.3µF to 90,000µF (was 1µF to 22,000µF).
- 10x faster capacitance measurement for large capacitors.
- Improved in-circuit accuracy.
- Improved user options system.
- New triple-slope measurement system vastly reduces the influence of parallel resistance and leakage current on capacitance readings.

Summary Specs

- ESR measurement from 0.00Ω to 40.0Ω. Can be used for simple low-ohms testing too.
- Audible alerts for different ESR levels.
- Automatic controlled discharge before analysis.
- Automatic and manual power-off.
- New backlit display.
- Very long battery life from a single AAA cell.

If you already have an ESR70 that accepts an AAA battery, contact us about having it upgraded to the new Gold spec.

It's only possible to show summary specifications here. Please ask if you'd like detailed data. Further information is also available on our website. Product price refunded if you're not happy.

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